

# DEVELOPMENT OF BROADBAND MICROWAVE COUPLER USING MICROSTRIP - SLOT LINES IN MIC ENVIRONMENT

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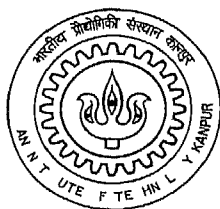
# **DEVELOPMENT OF BROADBAND MICROWAVE COUPLER USING MICROSTRIP-SLOT LINES IN MIC ENVIRONMENT**

*A Thesis Submitted  
in Partial Fulfillment of the Requirement  
for the Degree of*

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*by*

**SUKHBIR SINGH**



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**DEPARTMENT OF ELECTRICAL ENGINEERING  
INDIAN INSTITUTE OF TECHNOLOGY KANPUR**

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# CERTIFICATE

Submitted on 21/11/2000  
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It is certified that the work contained in the thesis entitled **Development of broad band microwave coupler using microstrip slot lines in MIC environment** by **Sukhbir Singh**, has been carried out under my supervision and that this work has not been submitted elsewhere for PGDIIT

  
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(8 November 2000

## Abstract

A novel method of realization of microstrip slot coupler with compensation is presented in this thesis. Generally microstrip couplers are realized using quarter wave planar transmission lines. This type of microwave coupler uses both sides of the substrate one side of which consists of microstrip line and the other side consists of a slotline. The substrate required for this coupler should have high permittivity due to dispersive nature of the slotline. Alignment for this type of coupler should be done in such a way that slotline must lie just beneath the microstrip line on the other side of the substrate. Analysis with and without compensation of the same coupler is made in this thesis by considering double symmetry of the reciprocal four port network. Compensation to the coupler is given by extending the length of slotline symmetrically on both sides of the terminating ends. Scattering parameters are derived for both the compensated and uncompensated cases of the coupler structure. The realization of the coupler with design parameters corresponding to 4.5 dB coupler on the alumina substrate with relative permittivity 9.8 and substrate thickness 0.635 mm is implemented over RT/Duroid substrate ( $\epsilon_r=10.2$ ) with thickness 0.325 mm. For this substrate approximate calculations are made using empirical formulae. Approximated theoretical center frequency is achieved around 6.7404 GHz. Measurements are taken over the band of interest (from 6.0 GHz to 8.0 GHz). The individual responses of the implemented coupler are measured and compared with the theoretical ones.

**To my parents**

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# CHAPTER 1

## 1.1 INTRODUCTION

Hybrids and couplers are indispensable components in rapidly growing applications of microwaves in electronics warfare radars and communication systems. The circuits are used frequently in frequency discriminators, balanced amplifiers, balanced mixers, automatic level control, and many other applications. A hybrid or a coupler can in principle be represented as a multi-port network. In the case of a four-port network, the four ports of structure are input, direct, coupled, and isolated. The two most important parameters that describe the performance of the network are its coupling factor and directivity.

A large number of papers on microwave directional couplers have appeared during the past 50 years. Some have reported advances in existing types, while others have described entirely new types or classes of directional couplers. The microwave Engineer is faced with a situation whereby he is frequently able to satisfy a specification by the use of one of several types of directional couplers. The choice is then made on the basis of cost, manufacturing techniques, and the physical layout, etc. Quite as often, the specification may be met only one type of coupler, so that knowledge of the entire field is desirable.

## 1.2 Classification of couplers

Attempts are often made to classify directional couplers in various ways. In Practice, it is found that no unique classification is possible. A classification may be based on one of many sets of conditions, and the particular sets in one scheme rarely coincide with those in another. Such schemes are given as following:

**1.2.1 Co- or Contra directional coupling between parallel Waveguides or Transmission Lines** In a co-directional coupler, the coupled wave travels in the

same direction as that of the wave in the main line while in the contra directional coupler the coupled wave travels in the opposite direction i.e. is coupled backward

**1 2 2 Interference or intrinsic coupling,** This applies to coupling between waveguide or transmission lines by means of longitudinal coupling (discrete or continuous) In an interference coupler the individual apertures or elemental parts of continuous aperture are not themselves directional in their coupling but the individual coupled waves tend to reinforce in one direction of propagation and to interfere destructively in the other In the case of intrinsic coupling the individual apertures are themselves directional in their coupling properties In this case the apertures may be cascaded to give an additional interference which enhances the directivity

**1 2 3 Classification by phase division** Most types of directional couplers may be classified as either 90 degree (quadrature) couplers or as 180 degree (magic Tee type) of couplers at least at their midband frequency They usually retain their 90degree or 180 degree property within some tolerance over the operating band but some asymmetric types do not

**1 2 4 Classification by the form of coupling** There are many ways in which coupling may be achieved some of the various methods are single aperture coupling multi aperture coupling continuous (longitudinal) aperture coupling branch line or guide coupling waveguide junctions ring couplers and quasi optical couplers

**1 2 5 Rectangular Waveguide** In the case of rectangular waveguide directional couplers the coupling is often classified as either narrow to narrow wall broad to broad wall or as narrow to broad wall The first two methods are most common and are called simple broad wall or narrow wall couplers respectively

This merely states that the types of guiding System in which the directional couplers may be designed e.g. rectangular ridged or circular waveguide coaxial and strip transmission lines quasi optical guides etc In general the fact is this that nearly all

Couplers may be analyzed in terms of the even and odd normal modes which means that the mathematical analysis of all these types are all much the same

We are interested in the implementation of directional coupler of which design involves planar transmission lines. Because this type of transmission structure is suitable in microwave integrated circuits. A planar geometry implies that the characteristics of the element can be determined from the dimensions in a single plane

### 1.3 Classification of transmission lines

Various forms of planar transmission lines have been developed for use in microwave integrated circuits [18]. The strip line, microstrip line, inverted micro strip line, slot line, co-planar waveguide, and co-planar stripline are respectively planar transmission lines. The circuits realized using any one of the aforementioned transmission lines or combinations of them have distinct advantages compared to conventional microwave circuits such as lightweight, small size, improved performance, better reliability, and reproducibility, and low cost. They are also compatible with solid state chip devices. Above mentioned planar transmission lines description is as follows

**1.3.1 Striplines** Strip line is one of the most commonly used transmission lines for passive MICs. The dominant mode in a stripline is TEM.

**Microstrip Lines** Unlike the stripline, the microstrip line is an inhomogeneous transmission line [15] since the field lines between the strip and ground plane are not contained entirely in the substrate. Therefore, the mode propagating along the microstrip is not purely TEM but quasi-TEM. The microstripline has become the most widely used transmission line medium because of its unique distinguishing quality needed by planar circuits. Its unique characteristics include a highly compact and rugged structure, Large scale integration feasibility, and ease of device and component implementation. To ensure single mode propagation, both strip width ( $w$ ) and substrate thickness ( $h$ ) are

usually held to a small fraction of the operating wavelength. At low frequency where quasi TEM propagation is assumed the microstrip is usually open structured. However at millimeter wave frequencies the microstrip line generally requires metallic shielding for radiation interference. And higher order mode excitation etc. There will be presence of waveguide modes in the shielded structure. To suppress waveguide mode propagation and allow only the quasi TEM mode to propagate along the line channel must be designed so that the cut off frequency of dominant waveguide mode which is generally hybrid mode as the currents in the strip flow both the longitudinal and transverse directions. Cut off of the dominant waveguide mode also depends on cross section of the shielding structure. In case of parallel coupled microstrips most of the E field lines between microstrips are parallel to ground plane whereas in odd mode case most of the E field lines between strips are perpendicular to ground plane. If the structure is pure TEM both odd and even mode phase velocities will be same. Here as strips are placed on dielectric substrate which causes different phase velocities in the even and odd modes. Comparatively in the even mode the fields will be more concentrated in the substrate than in the odd mode. In the odd mode case the field lines in the dielectric substrate will be lesser. Hence the effective dielectric constant is higher in the even mode case. Higher the effective dielectric constant causes lesser the phase velocity. By knowing the dispersion and impedance characteristic we can find the phase velocity and hence the directivity.

**1.3.3 Suspended and inverted Microstrip Lines** Suspended and inverted microstrip lines provide a higher Q (500 to 1500) than does a microstrip. The wide range of impedance values achievable makes these media particularly suitable for filters.

**1.3.4 Slotlines** The slotline configuration is useful in circuits requiring high

impedance lines series stubs short circuits and in hybrid combinations with microstrip circuits in microwave circuits This structure is not popular at millimeter wave frequencies The mode of propagation is non TEM and almost transverse (TE) in nature Various methods of analysis discussed in literature do not lead to the closed form expressions for slotline wavelength and impedance This becomes a serious handicap for circuit analysis and design especially when computer aided design techniques are used However closed form expression for characteristics impedance and slotline wavelength have been obtained by the numerically computed results

**1.3.5 Coplanar Lines** Coplanar waveguides are finding extensive applications in microwave integrated circuits Inclusion of coplanar waveguides in microwave circuits adds to the flexibility of circuit design and improves and improves the performance for some circuit functions Another promising configuration which is complementary to CPW is known as coplanar strips (CPS) both of these configurations belong to the category of coplanar lines

Wherein all conductors are in the same plane (i.e. on the top surface of dielectric substrate) A distinct advantage of these two lines lies in the fact that mounting lumped (active or passive) components in shunt or series configuration is much easier Drilling holes or slots through the substrate is not needed Coplanar waveguides and coplanar strips support quasi-TEM modes

## 1.4 Comparison of various MIC Transmission Media

For hybrid MIC applications microstrip slotline coplanar waveguide and coplanar strips have been used whereas for monolithic MICs microstrip has been used extensively It can generally be seen that cpw and cps combine some advantageous features of microstrip lines and slotlines Their power handling capabilities radiation loss Q

factors and dispersion behavior lie in between the corresponding values of microstrip and slotline. Perhaps the best feature of two coplanar lines is the ease of mounting components in series and shunt configurations, whereas microstrip lines are convenient only for series mounting and slotlines can accommodate only shunt mounted components.

## 1.5 Design of Directional Coupler

Microstrip directional couplers made of parallel coupled transmission lines exhibit a reduced directivity due to the difference in the phase velocity of the odd and even mode. To improve directivity, velocity equalization has been introduced by a capacitive loading of the odd mode using lumped capacitors at the ends of the coupled region.

It is the purpose of this thesis to apply this technique in the design of 3 dB hybrid so as to improve the directivity. Realization of this type of coupler can be done with a microstrip on the top of the substrate and slot line in the ground plane of the substrate. In this case, the length of the slotline between the two slot open circuits should be increased by a certain amount such that the additional length of slotline  $l_2$  requires capacitive loading. Let us assume that for an odd mode excitation at the ports 1&2 as determined mainly by the slotline is  $Z_{01}$  beneath the microstrip (length  $l_1$ ) and  $Z_{02}$  outside the microstrip (length  $l_2$ ). The electrical length of microstrip should be a quarter wavelength at the center frequency. The frequency response of the isolation depends on the ratio of  $l_2/Z_{02}$ . G. Schaller [6] and therefore for simplicity we designed the coupler with same slot width throughout.

Another possibility to compensate for the difference in phase velocity is to use slot-strip slot configuration described by Kohler [6]. In this case, the slot lines may be curved in order to obtain phase equalization. This is another interesting approach. Unfortunately, 3 dB coupling can not be realized in a planar microstrip microslot technique. The isolation is quite satisfactory.

which seems to prove the effectiveness of the curved slotlines (additional length  $\sim 1$  percent) for phase equalization

This microstrip slot coupler is a special class of symmetric couplers. It is particularly suitable for the realization of 3 dB couplers in MIC technology. De Ronde [2] originally proposed this coupler and Garcia [13] described an empirical design. In addition, Schiek [3] has made an analysis with the aid of the equivalent circuit of the branchline coupler. Several problems of practical microstrip-slot coupler design are treated in this thesis. It involves realization of 3 dB coupler in microwave integrated circuit on RT-Duroid (dielectric constant  $\epsilon_r = 10.2$ ). This thesis shows that the design data of the standard microstrip and slot line are good approximation of dimensioning the coupler cross section. In addition, design parameters that still remain to be determined, such as position of the reference planes, the definition of the slot-line characteristic impedance and the influence of the transmission line loss are investigated. The improvement of the coupler performance attainable through compensation is shown.

## **1.6 Organization of the thesis**

This thesis explores the possibility of realization of 3 dB broadband coupler using microstrip and slot lines in MIC environment. Which has compact size, lightweight and improved directivity. Hoffman [1] which the conventional directional couplers do not have. Chapter 1 presents survey of some important literature based on the couplers theory and their classification. Comparison of various transmission media and their types used in the MIC technology is also given in this chapter. The suggestions for development of broadband compensated coupler with microstrip and slot lines are considered here. Chapter 2 deals with microstrip slot coupler design theory and derivations of scattering parameters in both compensated and uncompensated form [1]. The scattering parameters involve input reflection coefficients which are derived using even-odd mode analysis of symmetrical structure of the coupler. Various characteristics of the microstrip and slot



lines are given in the description which are important for the design purpose of the microstrip slot coupler. The approach to the design of the coupler is discussed for both the compensated and uncompensated cases Hoffman [1]. Special cases of the microstrip slot coupler for ideal de Ronde [2], real uncompensated and real compensated are also discussed.

Chapter 3 deals with the practical design aspects of microstrip slot coupler Hoffman and Siegel [1] in supplementation of the theoretical analysis presented in chapter 2. Computation technique related to even-odd mode analysis is suggested here. Measured and theoretical results comparison is made for the implemented coupler and reference planes location is specified empirically. Finally the measured results are discussed. Chapter 4 presents summary, conclusions and scope for future work.

# CHAPTER 2

## Microstrip-Slot coupler design theory

### S-Parameters of Uncompensated and Compensated couplers

#### 2.1 INTRODUCTION

In this chapter description related to characteristic of microstrip line and slot line is given and scattering parameters of coupler having four port double symmetrical network are derived for both uncompensated and compensated cases using even odd mode analysis of coupler cross section. The uncompensated coupler and the coupler compensated by extending the slot lines are treated here and the approaches to design are also taking into account for the compensation of slot lines.

#### 2.2 Characteristics of microstrip line

Microstrip is an inhomogeneous transmission media. It consists of a dielectric substrate with a strip conductor on one surface and metallization to form a ground plane on the reverse surface.

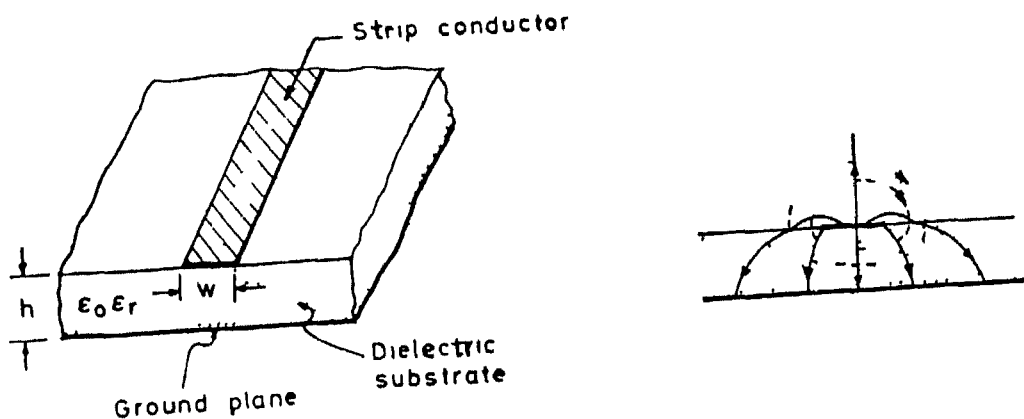


Figure 1.1 Microstrip line (quasi TEM mode)

Since the field lines between the strip and the ground plane are not contained entirely in the substrate. Therefore the dominant mode of propagation is quasi TEM that is the mode is dominantly TEM with small field components along the direction of propagation as shown in figure 1.1. The fields are confined to the vicinity of the strip conductor with a large concentration inside the dielectric substrate and less in the air region. Larger the value of  $\epsilon_r$  of the substrate the greater is the relative concentration of energy inside the substrate and lesser is the radiation. The characteristic impedance and guide wavelength of microstrip can be related to  $Z^a$  and  $\lambda$  in the following form  $Z = Z^a / \sqrt{\epsilon_{eff}}$  and

$\lambda = \lambda_0 / \sqrt{\epsilon_{eff}}$  [15] where  $\epsilon_{eff}$  is the effective dielectric constant of the microstrip medium. It is defined as the relative dielectric constant of an equivalent homogeneous microstrip which has the same phase velocity as the original (inhomogeneous) microstrip. The value of  $\epsilon_{eff}$  lies in the range  $0.5(1 + \epsilon_r) \leq \epsilon_{eff} \leq \epsilon_r$  depending on the value of  $w/h$ . For small strip widths ( $w/h \ll 1$ ) the fields are distributed nearly equally in the substrate and air regions and hence  $\epsilon_{eff}$  approaches the lower limit  $(1 + \epsilon_r)/2$ . For large strip widths ( $w/h \gg 1$ ) the electric field is confined mostly between the strip conductor and the ground plane. The microstrip then resembles a parallel plate capacitor with  $\epsilon_{eff}$  approaching  $\epsilon_r$ .

### 2.2.1 Effect of Dispersion

At higher frequencies the fields tend to concentrate more within the dielectric substrate. Consequently  $\epsilon_{eff}$  increases with an increase in frequency with the value approaching to the substrate permittivity  $\epsilon_r$  asymptotically in the limit frequency tends to infinity. The following closed form dispersion formula can be used to compute the frequency dependent  $\epsilon_{eff}$  (denoted as  $\epsilon_{eff}(f)$ ) in the range  $2 \leq \epsilon_r \leq 16$ ,  $0.06 \leq w/h \leq 16$  and  $f \leq 100\text{Hz}$  [15]

$$\epsilon_{eff}(f) = \left[ \left( \sqrt{\epsilon_r} \sqrt{\epsilon_{eff}} \right) / \left( (1 + 4F^{15}) + \sqrt{\epsilon_{eff}} \right) \right]^2 \quad 2.2.1(a)$$

Where

$$F = (4h \sqrt{(\epsilon_r - 1)} / \lambda) [0.5 + \{1 + 2 \log(1 + (w/h))\}^2] \quad 2.2.1(b)$$

$\epsilon_{eff}$  is the quasi-static value. Similarly, the effect of dispersion on  $Z$  can be obtained from the closed form expression [15] given in above equation

$$Z(f) = Z \left( \frac{\epsilon_{eff}(f) - 1}{\epsilon_{eff} - 1} \right) \sqrt{\epsilon_{eff} / \epsilon_{eff}(f)} \quad 2.2.1(c)$$

Where  $Z$  is the quasi-static value

## 2.2.2 Losses

The propagation loss (neglecting radiation) in a practical microstrip is primarily due to two types of dissipative losses: conductor loss due to the finite resistivity of the conductor and dielectric loss due to the finite loss tangent of the dielectric substrate. If we denote  $\alpha_c$  as the attenuation constant due to the conductor loss and  $\alpha_d$  as the attenuation due to the dielectric loss, both expressed in  $\alpha = (\alpha_c + \alpha_d)$  dB/unit length.

## 2.3 Slotline

The slotline consists of a slot etched from the conducting layer on one surface of a dielectric substrate with the opposite surface being bare.

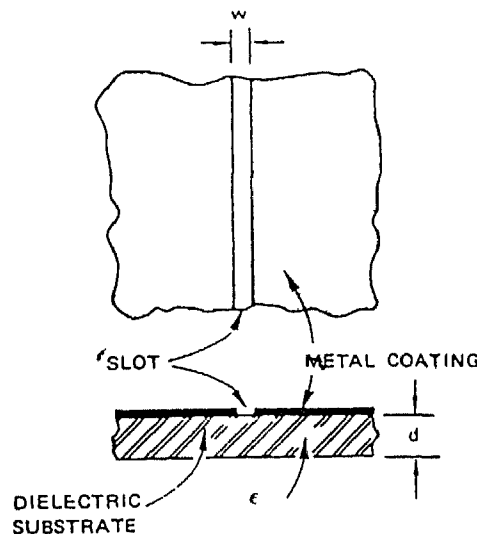


Figure 2.2 Slot line on a dielectric substrate

If substrate's permittivity is sufficiently high such as  $\epsilon_r = 10$  to  $30$  the slot mode wave length will be much smaller than the free space wave length and the fields will be closely confined near the slot. S B Cohn Slot line [10] is shown in the following figure

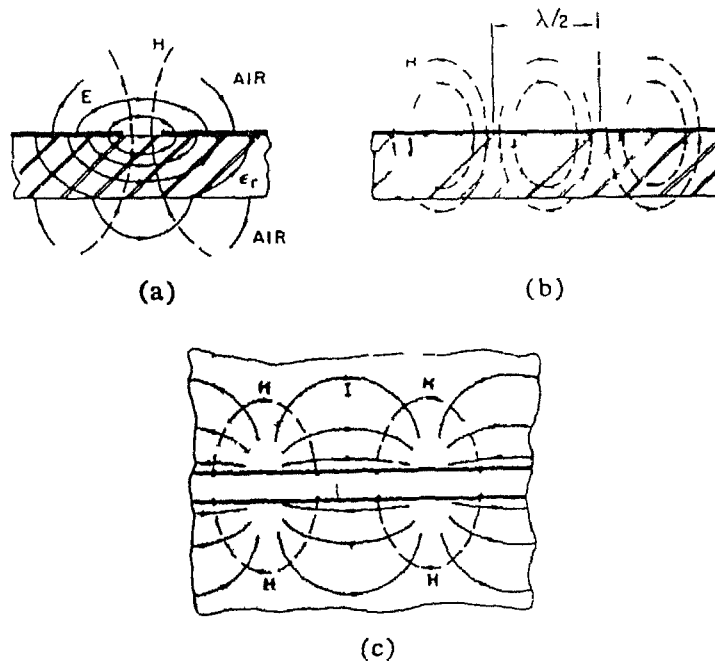


Figure 2.3 Field and current distribution (a) Field distribution in cross section (b) H field in longitudinal section (c) Current distribution on metal surface

The structure is thus complementary to that of microstrip. The electric field lines are oriented essentially across the slot whereas magnetic field lines have both transverse and longitudinal components. Figure 2.3(a) shows the slot-mode fields in a cross sectional view.

A voltage difference exists between the slot edges. The electric field extends across the slot; the magnetic field is perpendicular to the slot. Because the voltage occurs across the slot, the longitudinal view in figure 2.3 (b) shows that in air regions the magnetic field lines curve and return to the slot at half wavelength intervals. Consequently, a propagating wave has elliptically polarized regions that can be usefully applied in creating certain ferrite components. The current paths on the conducting surface are shown in figure 2.3(c).

The surface current density is greatest at the edges of the slot. A propagating wave has regions of elliptically polarized current and magnetic field in this view also. An interesting possibility for microwave integrated circuits is the use of slot lines on one side of substrate and microstrip lines on the other. When close to each other, coupling between two types of lines occurs; when sufficient far apart they will be independent. Coupling between a slot and a strip can be used intentionally in certain components. For example, parallel lengths of slot and strip can be made to act as directional coupler. The dominant mode of propagation is non TEM and hence the characteristic impedance and guide wavelength are dependent on frequency, although variation is slow. As compared to microstrip, slotline is more dispersive. The slotline mode resembles the  $TE_{10}$  mode of a rectangular waveguide but it differs from the waveguide in that it has no cutoff frequency.

Slot is a convenient medium for shunt mounting of discrete devices. It is particularly suitable for ferrite components that require regions of circularly polarized magnetic field. Its Q value is ( $\approx 100$ ).

## 2.4 Design approach

The microstrip slot coupler is particularly suitable for the realization of 3 dB couplers in MIC technology. The analysis is extended to include a lengthening of the slot line, which is used to compensate for the different phase velocities of the even and odd modes. There are design specifications reported for the compensated coupler. This type of coupler is realizable in planar Microstrip technique for a wide range of

comparing theoretical and measured electrical parameters of the coupler. Slot line is terminated in the circular form of open circuit so as to avoid sudden increase in impedance. It is basically done for impedance transformation from low to high in steps. Which provides discontinuity compensation [1]. As shown in following figure

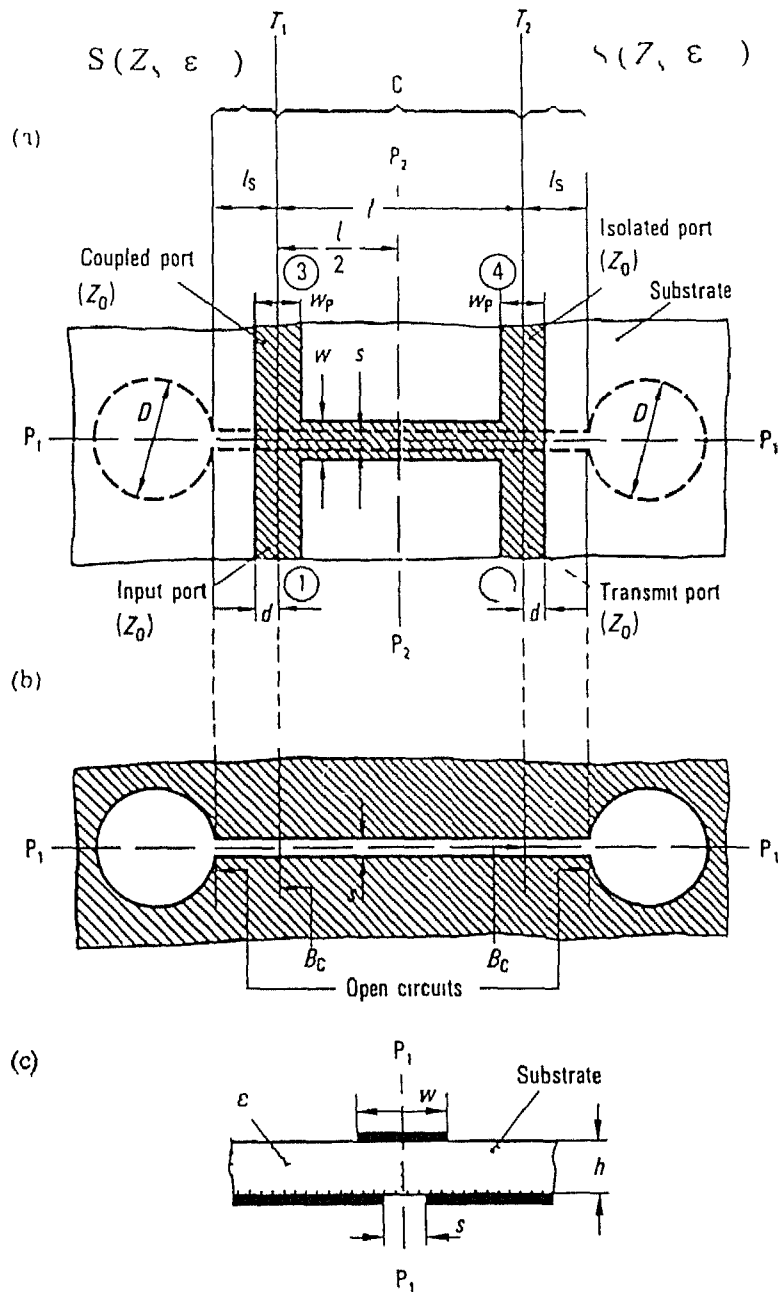


Figure 2.4 Configuration of Microstrip slot coupler (a) Upper side of substrate (b) Bottom side of substrate (c) Cross section

The conductors are assumed to be lossless and the open circuits at the ends of the slot line are assumed to be ideal. Also the coupler without compensation is defined by  $l_s = 0$ .

line are assumed to be ideal. Also the coupler without compensation is defined by  $l_s = 0$ . In this case, an ideal open circuit is located at the reference planes  $T_1$  and  $T_2$ . The hybrid has a double symmetry with respect to the planes  $p_1$  and  $p_2$  as shown in figure 2.5 and it is a reciprocal four-port network.

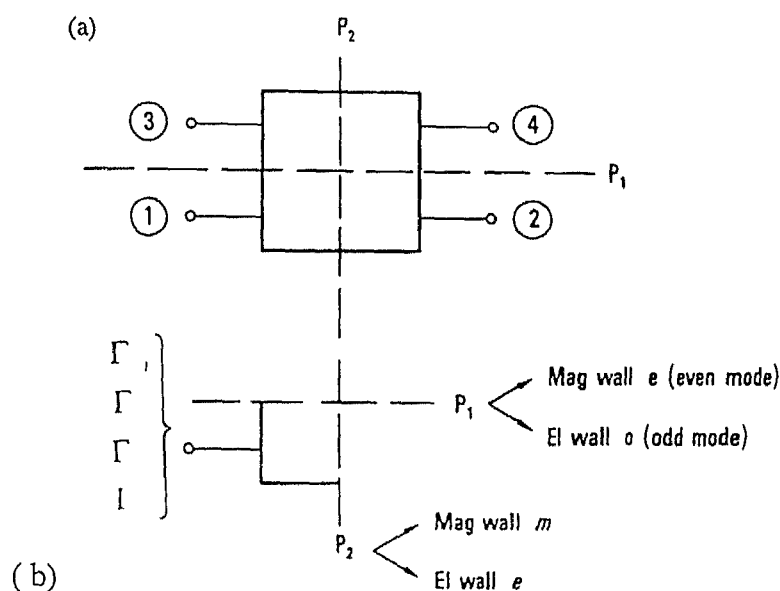


Figure 2.5 Common four-port network with double symmetry with respect to  $p_1$ ,  $p_2$  (a)

Configuration (b) Definition of reflection coefficients

The scattering parameters of the network can be written in terms of the even-odd mode reflection coefficients given below

$$S_{11} = \frac{1}{4} (\Gamma_{1e} + \Gamma_{1o} + \Gamma_{2e} + \Gamma_{2o}) \quad 2.4.1(a)$$

$$S_{21} = \frac{1}{4} (\Gamma_{1e} - \Gamma_{1o} + \Gamma_{2e} - \Gamma_{2o}) \quad 2.4.1(b)$$

$$S_{31} = \frac{1}{4} (\Gamma_{1e} + \Gamma_{1o} - \Gamma_{2e} - \Gamma_{2o}) \quad 2.4.1(c)$$

$$S_{41} = \frac{1}{4} (\Gamma_{1e} - \Gamma_{1o} - \Gamma_{2e} + \Gamma_{2o}) \quad 2.4.1(d)$$

The reflection coefficients  $\Gamma_{1e}$ ,  $\Gamma_{1o}$ ,  $\Gamma_{2e}$ , and  $\Gamma_{2o}$  are referenced to  $Z_0$  (the characteristic impedance of the connecting section) appearing at each end of the ports. On application



of certain combinations of perfect electric (pec) and perfect magnetic (pmc) planes at the planes of symmetry  $p_1$  and  $p_2$  for the computation the reflection coefficients from the original coupler configuration given in figure 2.4. The magnetic or electric walls likewise have to be applied to the symmetry planes  $p_1$   $p_2$ . The conductor pattern of the coupling section is fed at its ends from the terminals 1, 3 or 2, 4 of the coupler four port network as shown in figure 2.6(a). A magnetic wall at  $p_1$  corresponds to even mode excitation with equal terminal voltages  $V_1=V_2=V$ .  $p_1$  then divides the conductor pattern into two identical transmission lines each having the characteristic impedance  $Z_e$  and the effective permittivity  $\epsilon = (c/v)^2$  where  $c$  denotes the light velocity and  $v_e$  the phase velocity of the even mode. The relations are

$$Z_e = 2Z_M \quad \text{and} \quad \epsilon = \epsilon_e \quad 2.4.1(e)$$

Where  $Z_M$  is the characteristic impedance of the microstrip and  $\epsilon_e$   $\epsilon$  are the effective dielectric constants of the even mode of the coupled lines and the microstrip respectively.

An electric wall at  $p_1$  corresponding to odd mode excitation with opposite terminal voltages  $V_1=V_2=V$ . The wall divides the conductor pattern of the coupling section into two transmission lines. Each having the characteristic impedance  $Z_o$  and the effective permittivity  $\epsilon = (c/v)^2$  where  $v_o$  denotes the phase velocity of the odd mode.

A quasi slotline mode (quasi TE mode) appears along the entire line with the characteristic impedance  $Z_s$  as shown in figure 6(b).

$$Z_o = \frac{1}{2} Z_s, \quad \epsilon_o = \epsilon_s \quad 2.4.1(f)$$

Where  $Z_s$  and  $\epsilon_o$  are the characteristic impedance and the effective dielectric constant of the quasi slot line mode (quasi TE – mode) due the odd mode excitation.

As the structure is reciprocal passive having double symmetry with reference to the two symmetry planes  $P_1$  and  $P_2$ .  $Z_o$  is the characteristic impedance of the connecting lines.

In case of the even mode excitation  $Z_e=2Z_M$  and when mode of excitation is odd then

$$\text{characteristic impedance } Z_o = \frac{Z_s}{2}$$

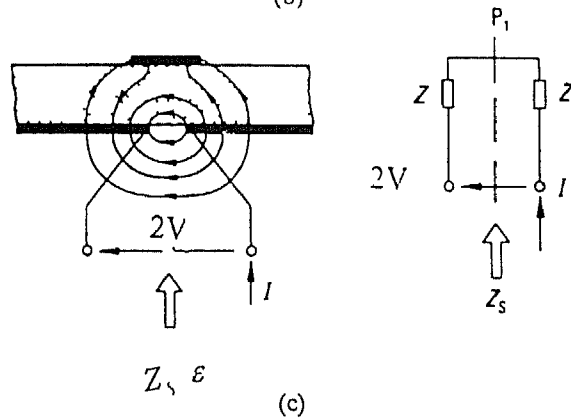
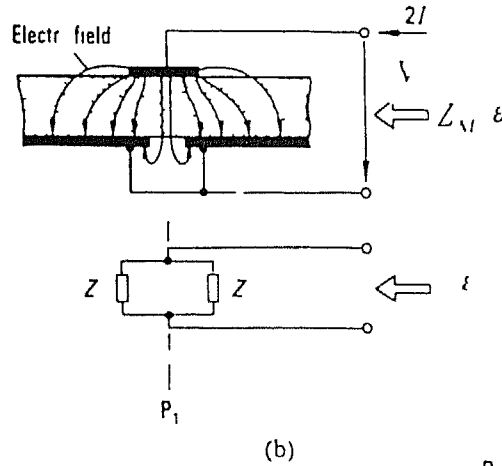
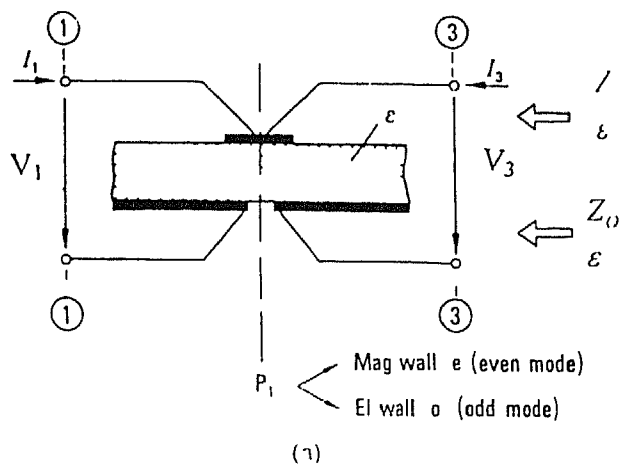


Figure 2.6 Even – odd mode excitation (a) Connection of terminals 1, 3 to coupling section (b) Even mode excitation with  $V_1 = V_2 = V$  Definition of  $Z_M$   $\epsilon_{ee}$  (microstrip mode) (c) Odd mode excitation with  $V_1 = V_2 = V$  definition of  $Z_S$   $\epsilon_{eo}$  (slot line mode)

microstrip length  $l = \theta_1 \epsilon \quad \lambda/4 = \theta_e$  slot length  $= \theta$

In microstrip mode boundary conditions are such that there is magnetic wall at  $p_1$  and magnetic/ electric wall at  $p_2$

First consider magnetic wall at  $p_2$  i.e. there is open circuit at  $p_2$  so input impedance seen at the junction of feed line and microstrip will be

$$Z = -jZ \cot\left(\frac{\theta}{2}\right) \quad 2.4.1 \text{ (I)}$$

We can derive expression for  $\Gamma$  at port 1 as follows

$$\Gamma = \frac{Z - Z}{Z + Z} \Rightarrow \frac{-jZ \cot\frac{\theta}{2} - Z}{-jZ \cot\frac{\theta}{2} + Z} \quad 2.4.1 \text{ (II)}$$

$$\Gamma = \frac{Z \cot\left(\frac{\theta}{2}\right) - jZ}{Z \cot\left(\frac{\theta}{2}\right) + jZ} \quad 2.4.1 \text{ (III)}$$

$$\begin{aligned} \Gamma &= \frac{\exp\left\{j \arctan\left[\frac{-Z}{Z \cot\left(\frac{\theta}{2}\right)}\right]\right\}}{\exp\left\{j \arctan\left[\frac{Z}{Z \cot\left(\frac{\theta}{2}\right)}\right]\right\}} \\ &= \exp\left\{-2j \arctan\left[\frac{Z}{Z} \tan\left(\frac{\theta}{2}\right)\right]\right\} \end{aligned} \quad 2.4.1 \text{ (IV)}$$

putting  $Z_c = 2Z_M$  in equation (IV)

$$\Gamma = \exp\left\{-j2 \arctan\left[\frac{Z_o}{2Z_M} \tan\left(\frac{\theta}{2}\right)\right]\right\} \quad 2.4.1 \text{ (g)}$$

Similarly we have derived expressions for  $\Gamma$ ,  $\Gamma_{on}$  and  $\Gamma_o$  given in equation 2.4.1(h) 2.4.1(i) and (j)

$$\Gamma = \exp\left\{j2 \arctan\left[\frac{Z}{2Z_M} \cot\left(\frac{\theta}{2}\right)\right]\right\} \quad 2.4.1 \text{ (h)}$$

$$\Gamma_{on} = \exp\left\{-j2 \arctan\left[\frac{2Z}{Z} \tan\left(\frac{\theta}{2}\right)\right]\right\} \quad 2.4.1 \text{ (i)}$$

$$\Gamma = \exp\{j2 \arctan[\frac{2Z}{Z_s} \cot(\frac{\theta}{2})]\} \quad 2.4.1(j)$$

Where

$$\theta = \omega l \sqrt{\epsilon} / c \quad (\text{Electrical length in case of even mode excitation})$$

$$\theta = \omega l \sqrt{\epsilon} / c \quad (\text{Electrical length in case of odd mode excitation})$$

The transmission parameters  $S_{ij}$  of uncompensated coupling section are in this way described as a function  $Z_M$ ,  $\epsilon$ ,  $Z_s$ ,  $\epsilon$ , length  $l$  and the angular frequency  $\omega$ . These equations apply for arbitrary values of  $Z_0$ ,  $Z_M$ ,  $Z_s$  and arbitrary phase velocities of  $v_e$  and  $v_o$ .

## 2.4.2 Even -odd Mode Analysis of the coupler with

**compensationLines** ( $l > 0$ ) Now considering the coupling section C shown by figure 2.4 which includes the slot compensation length  $l_s$ . Let S has the characteristic impedance  $Z_s$ , the effective permittivity  $\epsilon$  and the length  $l_s$ .  $Z_s$  and  $\epsilon$  differs from the parameters  $Z_s$  and  $\epsilon$  of the coupling section with odd mode excitation because in the case of the added slot length  $l_s$ , the strip conductor is missing from the other side of the substrate. With typical coupler on RT/Duroid substrate ( $Z_0 = 50 \Omega$ ,  $\epsilon_r = 10.2$ ),  $\epsilon_{eo} < \epsilon_{ee}$  in all cases, whence  $v_o > v_e$ . For  $l_s \ll \lambda/4$ , the slot length acts as a shunt capacitance  $c_c$  as shown in figure 2.7 that is only effective with odd mode excitation. This capacitance increases the transmission phase between the terminals 1, 3 and 2, 4 for odd mode excitation from  $\phi_o$  to  $\phi_o^*$ . For the selectable compensation frequency  $f_{co}$ , the compensation can be realized with equal values of  $\phi_o^*$  and the transmission phase for even mode excitation  $\phi_e = \theta_e = \omega l \sqrt{\epsilon_r} / c$  so that the directivity  $D(f_{co}) = 20 \log \left| \frac{S_{41}}{S_{31}} \right| = \infty$

and  $S_{11}(f_{co}) = 0$ . For even mode excitation  $B_C = \omega C_C$  is ineffective

Analysis is made by the similar method as adopted in analyzing the coupler without compensation, whereby it is assumed that susceptance

$$B_c = \omega C = \frac{1}{Z_s} \tan \theta_s$$

2 4 2 (a)

With  $\theta'_s = \omega l_s \sqrt{\epsilon} / c$  is shunted across the ends of the coupling section

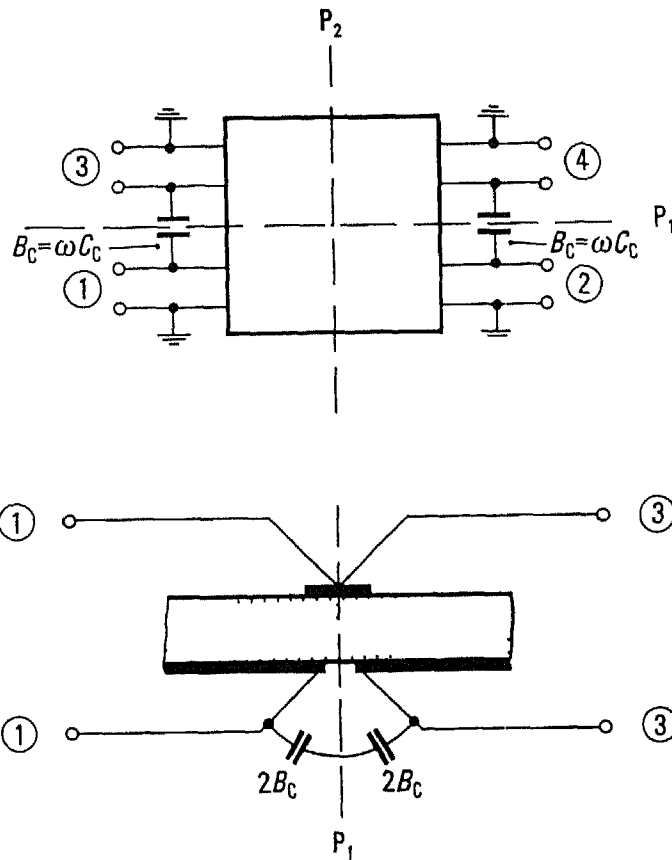


Figure 2 7 Coupler with compensation lines  $S(l_s > 0)$  represented by the input susceptance  $jB_c$

Again assume an ideal open circuit at the ends of the slot and an ideal junction at  $T_1$  and  $T_2$ . By applying appropriate magnetic and electric walls to the symmetry planes  $p_1$  and  $p_2$ , the reflection coefficients referenced to  $Z_0$  for odd mode excitation can be computed from the input susceptance of the transmission lines shunted with the susceptance  $2B_c$ .

For the slot mode (i.e. odd mode) compensation length  $l_s = \theta_s$

Then  $Z_{in \text{ effective}}$  can be computed by using lumped capacitors at the ends of coupled region. For that we assume an odd excitation at the ports 1 and 3 and the effective odd mode characteristic impedance can be determined mainly by the slotline impedance which is  $Z_{o1}$  beneath the microstrip (length  $l$ ) and  $Z_{o2}$  outside the microstrip (length  $l$ ).

Considering  $p_1$  as electric wall (corresponding to odd mode excitation) and  $p_2$  as magnetic wall for that

we can write

$$Z_{in \text{ effective}} = Z_{o1} + Z_{o2} \Rightarrow -j \frac{Z_s}{2} \cot\left(\frac{\theta}{2}\right) - j \frac{Z_s}{2} \cot(\theta_s) \quad 2.4.2(I)$$

For the above reflection coefficient at port 1 can be computed as following

$$\Gamma_{in} = \frac{Z_{eff} - Z}{Z_{eff} + Z} = \frac{\frac{Z_s}{2} \cot\left(\frac{\theta}{2}\right) + \frac{Z_s}{2} \cot(\theta_s) - jZ}{\frac{Z_s}{2} \cot\left(\frac{\theta}{2}\right) + \frac{Z_s}{2} \cot(\theta_s) + jZ} \quad 2.4.2(II)$$

$$\Gamma = \frac{\exp\left\{-j \arctan\left[\frac{Z}{\frac{Z_s}{2} \cot\left(\frac{\theta}{2}\right)} - \frac{Z}{\frac{Z_s}{2} \cot(\theta_s)}\right]\right\}}{\exp\left\{-j \arctan\left[\frac{Z}{\frac{Z_s}{2} \cot\left(\frac{\theta}{2}\right)} + \frac{Z}{\frac{Z_s}{2} \cot(\theta_s)}\right]\right\}} \quad 2.4.2(III)$$

$$\Gamma_{in} = \frac{\exp\left\{j \arctan\left[\frac{2Z_o}{Z_s} \tan\left(\frac{\theta}{2}\right) + \frac{2Z}{Z'_s} \tan(\theta'_s)\right]\right\}}{\exp\left\{-j \arctan\left[\frac{2Z}{Z_s} \tan\left(\frac{\theta}{2}\right) + \frac{2Z}{Z_s} \tan(\theta_s)\right]\right\}} \quad 2.4.2(IV)$$

From equation 2.4.2(a)  $B_C = \frac{1}{Z_s} \tan(\theta_s)$

Substituting this in equation 2.4.2(IV) we get

$$\Gamma_{om} = \exp \left\{ -j2 \arctan \left[ \frac{2Z}{Z_s} \tan\left(\frac{\theta}{2}\right) + 2B_c Z \right] \right\} \quad 2.4.2(a)$$

Similarly we have computed an expression for  $\Gamma_{oe}$

$$\Gamma_{oe} = \exp \left\{ -j2 \arctan \left[ -\frac{2Z}{Z_s} \cot\left(\frac{\theta}{2}\right) + 2B_c Z \right] \right\} \quad 2.4.2(b)$$

With  $\theta = \omega l \sqrt{\epsilon} / c$  and  $B_c$  according to equation 2.4.2(a) The parameters  $\Gamma$  and  $\Gamma$  can be computed with 2.4.1(g) and 2.4.1(h) because  $B_c$  is here ineffective. The scattering parameter  $S_{ij}$  of the coupler with  $l_s > 0$  are computed with 2.4.1(a) through 2.4.1(d). These equations apply for arbitrary values of  $Z_o$ ,  $Z_M$ ,  $Z_s$ ,  $\epsilon_{ee}$ ,  $\epsilon_{eo}$  and arbitrary parameters  $Z_s$ ,  $\epsilon_{eo}$ ,  $l_s$ .

## Special cases of Microstrip-Slot coupler

### 2.5.1 Ideal microstrip slot coupler

The ideal Microstrip Slot coupler can not be realized with the real coupler configuration shown in figure 2.4 because in this case  $\epsilon_{eo} < \epsilon_{ee}$ . It can however be used to derive design equations for the real coupler. The ideal coupler is characterized by the reflection coefficient  $S_{11}$  being zero and  $S_{41}$  being zero i.e. (infinite directivity) at all frequencies. Applying these conditions from 2.4.1(a) to 2.4.1(d) we can write

$$\Gamma_{ee} = \Gamma_{om} \text{ and } \Gamma_{em} = \Gamma_{oo} \quad 2.5.1(a)$$

For all frequencies

These equations apply only if two conditions with respect to the coupling section parameters exist. Condition one is that the coupling section has the same electrical length for the even and odd modes at all frequencies

i.e.  $\theta_e(f) = \theta_o(f) = \theta(f)$ . This results in

$$\epsilon_r = \epsilon_r = \epsilon \text{ and } \epsilon = \epsilon = \epsilon_e \quad 2.5.1(b)$$

and

$$Z_o = \sqrt{Z Z} = \sqrt{Z_M} Z_s \quad 2.5.1(c)$$

With these conditions satisfied we can use equations 2 4 1(b) 2 4 1(c) and 2 4 1(g) 2 4 1(j) for the transmission coefficient  $S_{21}$  and the coupling coefficient  $S_{31}$  assuming the following simplified form

$$S_{21}(f) = \frac{[1 - C^2]^{1/2}}{[1 - C^2]^{1/2} \cos \theta + j \sin \theta} \quad 2 5 1 (d)$$

$$S_{31}(f) = \frac{jC \sin \theta}{\sqrt{1 - C^2} \cos \theta + j \sin \theta} \quad 2 5 1 (e)$$

Where

$$\theta = \frac{2\pi}{\lambda} \sqrt{\epsilon} l$$

At the center frequency  $\theta = \frac{\pi}{2}$  These are the parameters of an ideal TEM coupler

where

$$C = \frac{Z_0 - Z_o}{Z_0 + Z_o} \quad 2 5 1 (f)$$

Which gives

$$C = \frac{4Z_M - Z_S}{4Z_M + Z_S} \quad 2 5 1 (g)$$

Where  $C$  is the coupling coefficient at the center frequency  $f_c = \frac{c}{4l\sqrt{C}}$   $l = \frac{\lambda}{4}$  where

$\lambda$  is the line wavelength The magnitude of  $S_{21}(f_c) = -j\sqrt{1 - C^2}$  is here at minimum and that of  $S_{31}(f_c) = C$  is at maximum For a given center frequency coupling loss  $a_c = -20\log C$  for a given characteristic impedance  $Z_0$

Combination of equations 2 5 1 (c) 2 5 1 (f) and 2 5 1 (g) yield the synthesis equations

$$Z_M = \frac{Z}{2} \sqrt{\frac{1+C}{1-C}} \quad 2 5 1 (h)$$

$$Z_S = 2 Z_0 \sqrt{\frac{1-C}{1+C}} = \frac{Z_0^2}{Z_M} \quad 2 5 1 (i)$$

For a 3 dB coupler in a  $50 \Omega$  the characteristic impedances



$Z_M=60.35 \Omega$  and  $Z_S=41.4 \Omega$  are obtained

At this point it is important to consider the differences between the present analysis of the microstrip slot coupler and the hybrid branchline coupler analysis. In the latter the microstrip slot coupler is treated as a special case of the  $\pi$  type hybrid branch line coupler as shown in figure 2.8 (a). It consists of two identical parallel lines  $G_p$  and a series transmission line  $G_s$  each having the length  $l$ . To allow comparison of the two analyses the hybrid branch line coupler is subjected to even mode excitation (magnetic wall at  $p_1$  according to figure 2.8(b)) for identical phase velocities in all lines. It follows from this that  $Z_e=Z_p$  for even mode excitation and  $Z_o$  is represented by  $Z_p$  in

parallel to  $Z/2$  resulting in 
$$Z = \frac{Z_p Z}{2Z_l + Z_s} \quad 2.5.1(i)$$

For odd mode excitation. This results in the synthesis equations  $Z_p=Z_o \sqrt{\frac{(1+C)}{(1-C)}}$  and

$$Z_{se} = Z \sqrt{\frac{1-C^2}{C}}$$

for the hybrid branchline coupler which yield  $Z_p=120.7\Omega$  and  $Z_{se}=50\Omega$  for a 3 dB coupler with a  $Z_o$  of  $50\Omega$

The analysis of this thesis and that of Schiek's paper agree if the microstrip mode-excited coupling section of figure 2.6(b) with the characteristic impedance  $Z_M$  is modeled by a parallel combination of two impedances  $Z_p$  and the slot mode-excited coupling section of figure 2.6(c) with the characteristic impedance  $Z_S$  is modeled by a parallel combination of  $Z_{se}$  and  $2Z_p$ . Applied to the above mentioned 3 dB coupler with  $Z_o=50 \Omega$  a parallel combination of two impedances  $Z_p=120.7 \Omega$  results in impedance of  $60.35 \Omega$  which corresponds to  $Z_M$  according to 2.5.1(h) and a parallel combination of  $Z_{se}=50 \Omega$  and  $2Z_p=241.4 \Omega$  results in an impedance of  $41.4 \Omega$  which corresponds to  $Z_S$  according to 2.5.1(h).

A general rigorous analysis of the coupling section such as a numerical analysis from

the field quantities permits – basically – only the calculation of the even –and odd – mode impedances  $Z_e$  and  $Z_o$ . Since in Schiek's equivalent circuit  $Z_o$  is modeled by a combination of both  $Z_p$  and  $Z_{se}$  it is not possible to calculate  $Z_{se}$  separately from the odd mode excited coupling section i.e. from the field quantities

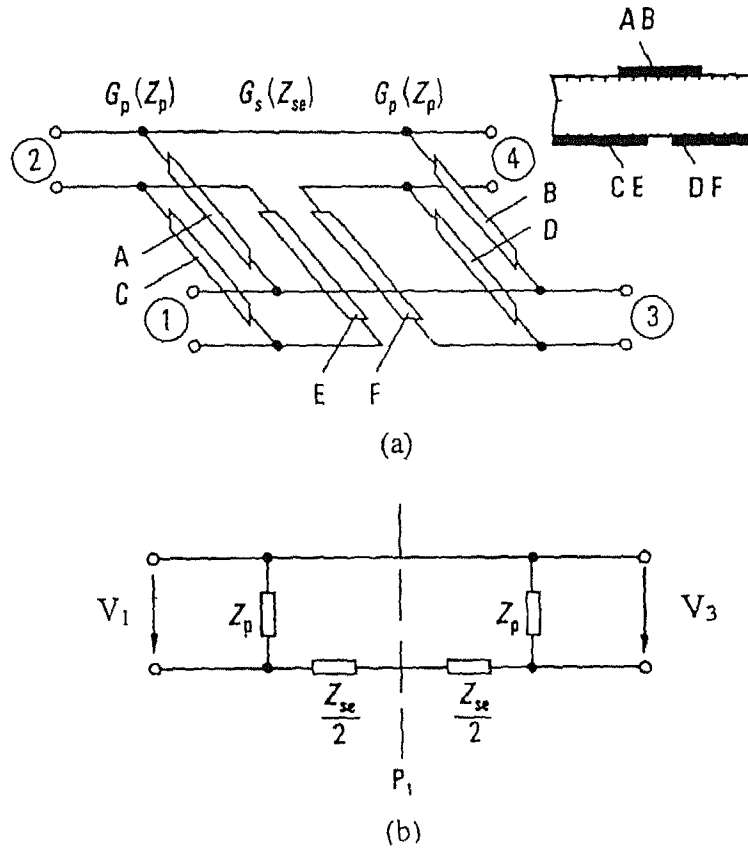


Figure 2.8 Comparison with hybrid branchline coupler analysis (a) Equivalent diagram of hybrid branch line coupler (b) Even-odd mode excitation of (a)

Therefore the interpretation of  $Z_{se}$  as the isolated slot line impedance which has been adopted in the ideal coupler where a  $50\ \Omega$  slot line impedance is given for the 3 dB coupler with  $Z_o = 50\ \Omega$  is strictly speaking an approximation

However the characteristic impedances  $Z_M$  and  $Z_S$  of the analysis of this thesis are defined directly from the field quantities and therefore characterize the coupling section

exactly

## 2 5 2 Real Uncompensated Coupler

In a real uncompensated coupler  $l_c = 0$  as shown in Figure 2 4 Assuming commercial couplers of conventional dimensions  $\epsilon_{eo} \leq \epsilon_{ee}$  This gives the odd -mode phase velocity greater than the even mode phase velocity However the following is always satisfied

$$2 \left\{ \frac{\epsilon - \epsilon_e}{\epsilon_e + \epsilon} \right\} \leq 1 \quad 2 5 2 (a)$$

Equation 2 5 2(a) implies that for feeding at port 1 the direct port 2 the coupled port 3 and the isolated port 4 remains unchanged The coupler parameters are no longer ideal especially its directivity which is  $D < \infty$  and  $|S_{11}| > 0$  However  $|S_{11}|$   $|S_{41}| \ll 1$  remain if the matching condition 2 5 1(c) is taken in to account or the synthesis equations 2 5 1(h) and 2 5 1(i) of the ideal coupler are used for designing The scattering parameters are computed with equations 2 4 1(a)–2 4 1(d) and 2 4 1(g) 2 4 1(j ) whereby  $S_{21}(f)$  and  $S_{31}(f)$  closely approximate the ideal coupler parameters computed with 2 5 1(d) and 2 5 1(e) For the center frequency  $f_c = \frac{c}{4L\sqrt{\frac{\epsilon + \epsilon_e}{2}}}$  the

reflection coefficient and directivity can be obtained from equations 2 4 1(a)–2 4 1(d) and 2 4 1(g) 2 4 1(j) by the first order approximation

$$S_{11} \ S_{41} \ll 1$$

And a first order approximation yields

$$S_{11}(f_c) = j \frac{\pi}{4} \left\{ \frac{\epsilon - \epsilon_e}{\epsilon_e + \epsilon} \right\} C \sqrt{1 - C^2} \quad 2 5 2 (b)$$

$$D(f_c) = 20 \log \left| \frac{\pi}{4} \left\{ \frac{\epsilon - \epsilon_e}{\epsilon_e + \epsilon} \right\} \frac{1 - C^2}{C} \right| \quad 2 5 2 (c)$$

Where  $f_c$  is the center frequency

The essential performance resembles that of the microstrip coupler

## 2 5 3 Real compensated Coupler

For real couplers with  $\epsilon_{co} < \epsilon_{ee}$  matching and compensation i.e.  $S_{11}=0$   $S_{41}=0$  can be realized with the aid of the compensation lines S only at a single arbitrarily chosen frequency Viz the compensation frequency  $f_{co}$ . For this it is essential that

- 1) The matching condition 2 5 1 (c) be satisfied
- 2)  $Z'_s$  and  $l_s$  be approximately chosen and
- 3) That  $Z_s$  is increased slightly to  $Z_s^*$

To derive the compensation conditions we refer to the section entitled Ideal Microstrip -Slot coupler according to which ideal coupler behaves at  $f_{co}$  can be obtained if matching condition 2 5 1(c) is satisfied and  $\epsilon_e$  is equal to  $\epsilon$ . For the real coupler the matching condition can be easily satisfied by appropriate choice of the characteristic impedance  $Z_M$  and  $Z_s$  and according to the synthesis equation 2 5 1(h) and 2 5 1(i) but still  $\epsilon < \epsilon_{ee}$  because of the difference in field distributions of the Microstrip and slot modes. At one particular frequency however namely at the compensation  $f_{co}$  the effect of difference in  $\epsilon_{co}$  and  $\epsilon_{ee}$  can be compensated if we extend the slotline by a certain amount  $l_s$ . The additional slot lines S resulting from this extension leave the transmission characteristics of the Microstrip mode -excited coupling section unchanged because here the voltage across the slot of coupling is zero. These additional lines S modify the transmission characteristics of the slot mode excited coupling section only.

This real slot mode excited coupling section with compensation lines hereinafter referred (R) and is shown in figure 2 9(a)

To obtain compensation at  $f_{co}$  the transmission phase of (R) applying to  $l_s=0$  has to be increased by adding capacitive end loading. This is accomplished by lengthening the slot by certain amount  $l_s$  at which the transmission phase reaches to  $2\pi l f_{co} \sqrt{\epsilon} / c$  of the Microstrip mode excited coupling section. Unfortunately the capacitive end loading decreases the input impedance level of (R)

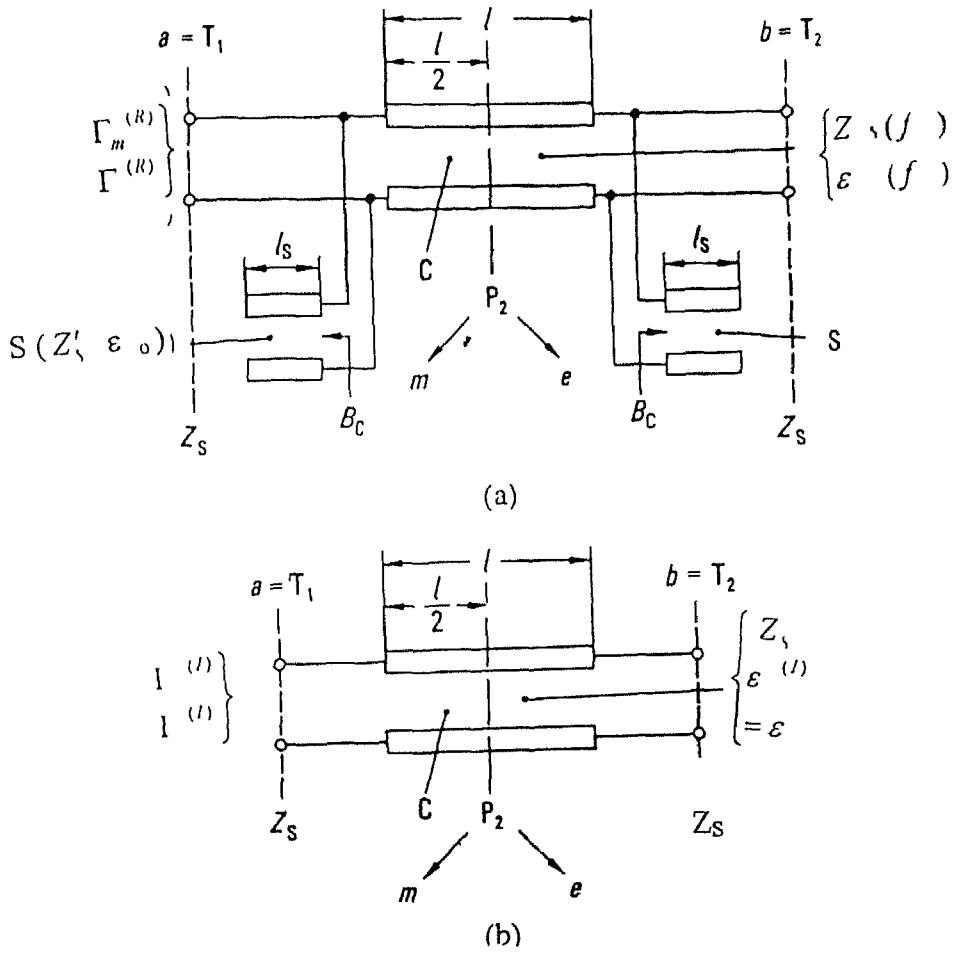


Figure 2.9 For synthesis the compensation conditions (a) Real coupling section (R) with compensation lines S excited with slot line mode (b) Fictitious ideal coupling section (I) with  $\epsilon_{eo}^{(I)} = \epsilon_{ee}$  excited with slot line mode

To compensate for this the characteristic impedance  $Z_s$  has to be increased to  $Z_s^*$  equation for  $l_s$  and  $Z_s^*$  which the S parameters  $S_{aa}^{(R)}$  and  $S_{ab}^{(R)}$  of (R) have at  $f_{co}$ , Now assume the characteristic impedance of lines S to be  $Z_s \cong Z_s^*$

To derive equations for  $l_s$  and  $Z_s$  the S parameters  $S_{aa}^{(I)}$  and  $S_{ab}^{(I)}$  of an ideal slot mode excited coupling section without compensation lines as shown in figure 2.9(b)

hereinafter referred to as (I) with the parameters  $Z_s$  according to

2.5.1 (i)  $\epsilon^{(I)} = \epsilon^{(I)}$  would provide compensation at any frequency

Therefore the compensation conditions are  $S_{aa}^{(I)} = S_{aa}^{(I)}$  and  $S_{ab}^{(I)} = S_{ab}^{(I)}$  at  $f_{co}$  all parameters being referred to as  $Z$ . The terminals a and b are the ends  $T_1$  and  $T_2$  of all the coupling section. Again Utilizing the symmetry of the configurations (I) and (I) we can calculate the S parameters  $S_{aa}$ ,  $S_{ab}$  from one port reflection coefficients  $\Gamma$  and  $\Gamma$  as we did for the whole coupler in 2.4.1(g) 2.4.1(j). The above mentioned compensation conditions can then be divided into two alternative compensation conditions

$$\Gamma_m^{(I)}(f_{co}) = \Gamma_m^{(I)}(f_{co}) \quad 2.5.3(a)$$

$$\Gamma_e^{(I)}(f_{co}) = \Gamma_e^{(I)}(f_{co}) \quad 2.5.3(b)$$

Where  $\Gamma_m^{(I)}$  and  $\Gamma_e^{(I)}$  are the input reflection coefficients referenced to  $Z_s$  of the ideal configuration as magnetic wall or electric wall is applied to  $p_2$ . The parameters  $\Gamma_m^{(I)}$  and  $\Gamma_e^{(I)}$  are the corresponding reflection coefficients likewise referenced to  $Z_s$  of the real configuration. equations 2.5.3(a) and 2.5.3(b) can only be satisfied by altering the characteristic impedance of the slotline

$$f_{co} = f_M = \frac{c}{4l\sqrt{\epsilon}} \quad 2.5.3(c)$$

Further assume for simplicity that slight increase in  $Z_s$  to  $Z_s^*$  required for compensation will not alter the effective permittivity  $\epsilon$  i.e.  $\epsilon_{eo} = \epsilon_{eo}$ . Thus 2.6(a) and 2.6(b) will yield the compensation conditions

$$2\arctan\left[B - \frac{Z_s}{Z_s^*} \cot\left(\frac{\pi lf_M \sqrt{\epsilon}}{c}\right)\right] = \frac{\pi}{2} \quad 2.5.3(d)$$

$$2\arctan\left[B + \frac{Z_s}{Z_s} \tan\left(\frac{\pi lf_M \sqrt{\epsilon}}{c}\right)\right] = \frac{\pi}{2} \quad 2.5.3(e)$$

With the input susceptance (obtaining at the compensation  $f_{co} = f_M$ )

$$B_{c_{co}} = 2\pi f_M C_{c_{co}} = \frac{1}{Z_s} \tan\left(\frac{2\pi l_s f_M \sqrt{\epsilon}}{c}\right) \quad 2.5.3(f)$$

Of the compensation lines S In 2.5.3(f) it is assumed that the characteristic impedance of these lines is  $Z_s = Z_s$  and their effective permittivity is  $\epsilon_{co} = \epsilon_{co}$ . Equations 2.5.3(d) and 2.5.3(e) yield after rearrangement the increased characteristic required for compensation at the compensation frequency  $f_{co} = f_M$

$$Z_s^* = \frac{Z_s}{2} \left[ \cot\left(\frac{\pi}{4} \sqrt{\frac{\epsilon}{\epsilon}}\right) + \tan\left(\frac{\pi}{4} \sqrt{\frac{\epsilon}{\epsilon}}\right) \right] \quad 2.5.3(g)$$

The length of the compensation lines which are assumed to have approximately the parameters  $Z_s = Z_s^*$  and  $\epsilon_{co} = \epsilon_{co}^* = \epsilon_{co}$  should be chosen according to

$$l_s = l \sqrt{\frac{\epsilon_c}{\epsilon}} \frac{2}{\pi} \arctan\left\{ \frac{Z_s}{Z_s} \left[ 1 - 2 \sin^2\left(\frac{\pi}{4} \sqrt{\frac{\epsilon}{\epsilon}}\right) \right] \right\} \quad 2.5.3(h)$$

As an alternative to extending the slot line by  $l_s$  it is possible to shunt both ends of the coupling section at  $T_1$   $T_2$  with a capacitance

$$C_{c_{co} = \epsilon_0} \frac{\eta}{Z_s} \frac{2}{\pi} l \sqrt{\epsilon} \left[ 1 - 2 \sin^2\left(\frac{\pi}{4} \sqrt{\frac{\epsilon}{\epsilon}}\right) \right] \quad 2.5.3(i)$$

Parallel to the slot line  $\eta_0 = 120\pi$  and  $\epsilon_0 = \frac{1}{(3.6\pi)} pF/cm$ . The values of  $l_s$  according to

2.5.3(h) agree to within a few percent with the first - order approximation however does not account for the increase in the characteristic impedance  $Z_s$  according to 2.5.3(g)

# CHAPTER 3

## Practical Design Aspects of Microstrip- Slot coupler

**3 1 Introduction** Practical designing aspects of microstrip slot coupler on an RT/ Duroid substrate (  $\epsilon_r=10.2$  ) are treated in supplementation of the theoretical analysis of the coupler presented in chapter 2. Comparison with implemented coupler yields rules for specification of the reference planes at the ends of coupling section and for the appropriate choice of the definition for the slot line characteristic impedance

Specifications of the standard versions of 4.5 dB microstrip – slot coupler

Slot width  $s = 0.14$  mm

Microstrip width =  $0.42$  mm

Substrate thickness  $h=0.635$  mm

$D= 7$  mm

$r= 5$  mm

$b= 0.5$  mm

Feed line width =  $0.62$  mm

$\epsilon_r=9.8$

Microstrip length =  $5.62$  mm

Compensation slot length  $l_s=0.81$  mm

To supplement the theoretical analysis presented in chapter 2 several problems of practical Microstrip slot coupler design are treated here. Emphasis is given to the realization of 3 dB couplers in microwave integrated circuit technology on RT/Duroid substrate (dielectric constant=10.2). The configuration of this coupler is



given in figure 3 1 This chapter shows that the design data for the standard microstrip and slot line are good approximations for dimensioning the coupler cross section The accuracy of the analysis in chapter2 is verified by a comparison with the measurement data

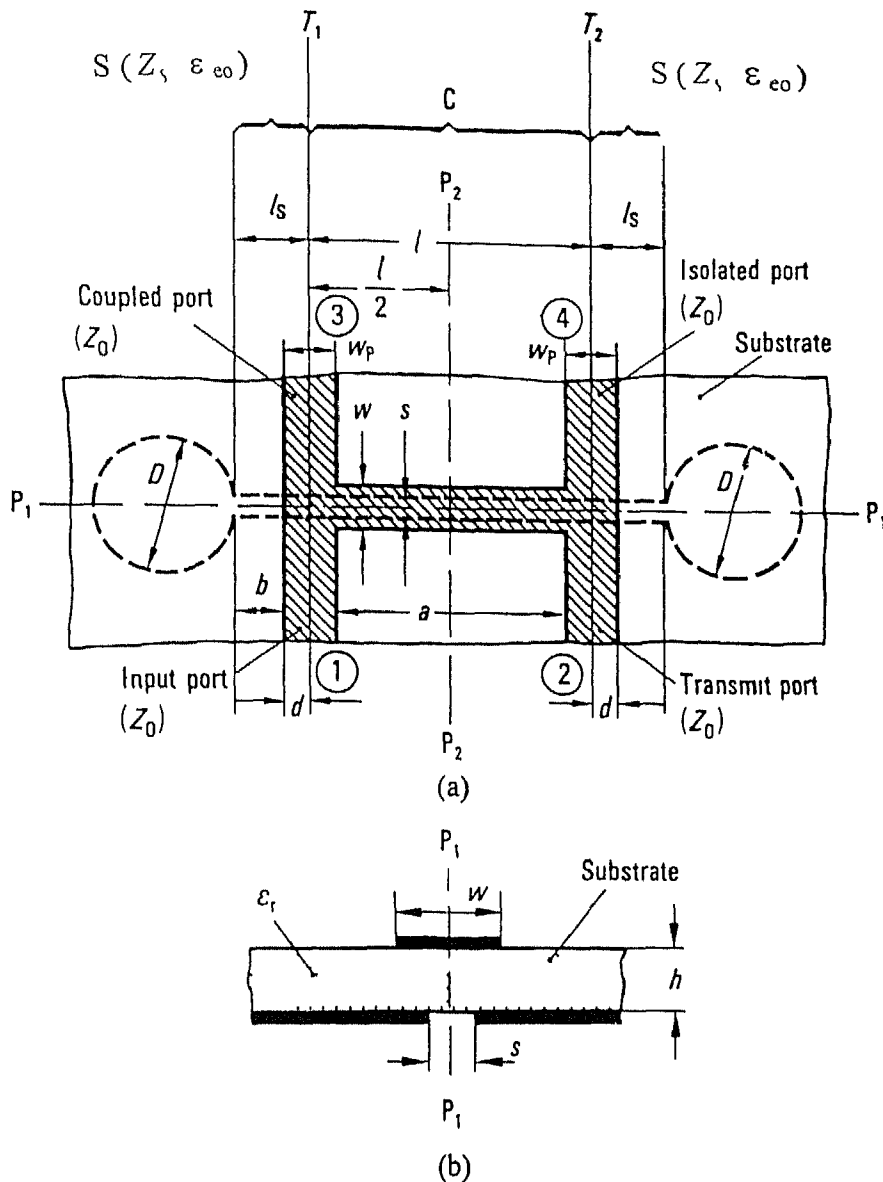


Figure 3 1 Configuration of the microstrip -slot coupler (a) Upper side of the substrate (b) Cross section

In addition design parameters that will remain to be determined such as position of the reference planes The definition of the slot line characteristic impedance and the

influence of the transmission line loss are investigated. The improvement of the coupler performance attainable through compensation is shown with the help of theoretical analysis.

## 3.2 Simplified computation of even and odd mode Parameters

The accurate comparison of the characteristic impedance  $Z_M$  and effective permittivity  $\epsilon_{ee}$  for the microstrip mode (even mode excitation) and the corresponding values  $Z_S$  and  $\epsilon_{eo}$  for the slot mode (odd mode excitation) of the original coupler cross section shown in figure 3.1 requires an extensive numerical analysis. In practice, such data are often not available. Analyses are however usually available for the standard microstrip transmission line. These analyses can provide useful approximate parameter values for the microstrip slot configuration. The errors in  $Z_M$  and  $\epsilon_{ee}$  under the assumption that the slot width  $s$  is negligibly small for  $s/h < 0.3$ . Coupling coefficient of  $\leq 6$  dB is taken into account here. Thus the parameters for 3 dB microstrip – slot couplers for the conventional microwave frequency range (approximately 2–18 GHz) are covered by the approximations of the standard microstrip and slot line.

## 3.3 Comparison of measured and theoretical Results

### 3.3.1 Implemented coupler

Microstrip slot coupler with a center frequency  $f_c$  in the region of 6.74024 GHz is implemented on RT/Duroid substrate using thin film technology with  $6 \mu m$  thick copper conductors. View of which is shown in figure 3.2.

The coupler is partially compensated. This means that the length  $l_s$  of the compensation slot line is  $> 0$ . But the compensation condition 2.5.3(h) of chapter 2 for  $l_s$  is not met, the coupler is said to be 3 dB coupler if it satisfies the matching condition  $Z_0 = \sqrt{Z_e Z} = \sqrt{Z_M Z_S}$ . But this type of coupler is not realizable because

of non-availability of correct thickness substrate, mask aligning technique. So the design corresponding to the available substrate is actually for 5.31- dB coupler with slot width 0.14 mm and center frequency 6.74024 GHZ.

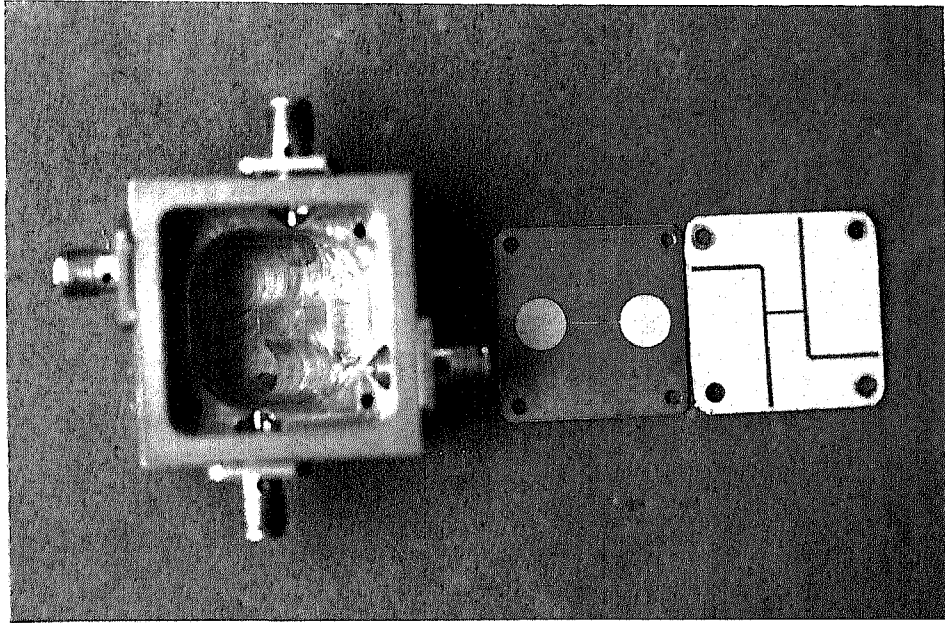


Figure 3.2 Photograph of the unassembled view of the microstrip-slot coupler

The measured frequency responses of the transmission loss  $a_{21} = -20\log|S_{21}|$ , coupling loss  $a_{31} = -20\log|S_{31}|$  and isolation loss  $a_{41} = -20\log|S_{41}|$  of the implemented coupler are compared with the theoretical responses.  $S_{ij}$  are the scattering parameters of the coupler. The measured frequency responses of  $a_{21}$  and  $a_{31}$  refer the whole circuit on the substrate shown in figure 3.1, and include the attenuation of 0.1 dB in all two feeding lines. The theoretical values of  $S_{21}$ ,  $S_{31}$  were compared with the aid of the analysis given in chapter 2, with  $Z_0 = 50\Omega$

### 3.3.2 Empirical Determination of Reference Plane Location and of proper slot Line Impedance Definition

For the calculation of the theoretical responses of  $a_{21}(f)$  and  $a_{31}(f)$  for the coupler design in general two questions arise First at what position should the reference planes  $T_1$   $T_2$  be placed? And second which definition should be chosen for the computing the characteristic impedance  $Z_S$  and  $Z_M$ ? The answers can be obtained by comparing the theoretical responses with the measured frequency responses

For the first question the tentative positions of the reference planes described by the distance  $d$  between  $T_1$  (or  $T_2$ ) and outer edge of the conductor of the feeding microstrip line is  $d = \frac{w_p}{2}$  and the corresponding value of the effective length  $l$  of the coupling section is shown

Here we have considering  $d = \frac{w_l}{2}$  for the calculation center frequency corresponding to which  $l = \frac{\lambda}{4} = 5.62 \text{ mm}$

The given substrate thickness is 0.325 mm and  $\epsilon_r = 10.2$  on which coupler fabricated consist microstrip line width  $w = 0.42 \text{ mm}$  and slot line width  $s = 0.14 \text{ mm}$  For microstrip line  $\frac{w}{h} = 1.292$  and for slot case  $\frac{s}{h} = 0.43076$

### 3.4 Approximate Calculations

Closed form expressions for characteristic impedance and slotline wavelength have been obtained by curve fitting the results based on the Cohn's analysis These expressions have an accuracy of about 2% for the following parameters

$$9.7 \leq \epsilon_r \leq 20$$

$$0.02 \leq \frac{w}{h} \leq 1.0 \text{ and } 0.01 \leq \frac{h}{\lambda_0} \leq \left( \frac{h}{\lambda} \right)$$

Where  $\left( \frac{h}{\lambda} \right)$  is the cut off value for the  $TE_{10}$  surface wave mode on slotline and is

$$\text{given by } \left( \frac{h}{\lambda} \right) = \frac{0.25}{\sqrt{\epsilon_r - 1}}$$

For  $0.2 \leq \frac{s}{h} \leq 1.0$

$$\frac{\lambda_s}{\lambda} = 0.987 - 0.21 \ln \varepsilon_r + \frac{s}{h} (0.111 - 0.0022 \varepsilon_r) - \left( 0.053 + 0.41 \frac{s}{h} - 0.0014 \varepsilon_r \right) \ln \left( \frac{h}{\lambda_0} 10^9 \right) \quad 3.4(a)$$

By substituting the value of  $\varepsilon_r$  and  $\frac{s}{h}$  in above equation we get

$$\frac{\lambda_s}{\lambda} = 0.4794$$

And  $\varepsilon_{co} = 5.711$

For  $0.2 \leq \frac{s}{h} \leq 1.0$

$$Z_s = 113.1923 - 257 \ln \varepsilon_r + 1.25 \frac{w}{h} (114.59 - 22.531 \ln \varepsilon_r) + 20 \left( \frac{w}{h} - 0.2 \right) \left( 1 - \frac{w}{h} \right) \left[ 0.15 + 0.1 \ln \varepsilon_r + \frac{w}{h} (-0.79 + 0.899 \ln \varepsilon_r) \right] \left\{ \left[ 10.25 - 2.171 \ln \varepsilon_r + \frac{w}{h} (2.1 - 0.617 \ln \varepsilon_r) - \frac{h}{\lambda} 10^2 \right]^2 \right\} \quad 3.4(b)$$

Putting the values of  $\varepsilon_r = 10.2$ ,  $\frac{s}{h} = 43074$  and  $\frac{h}{\lambda} = \left( \frac{h}{\lambda} \right) = \frac{0.25}{\sqrt{\varepsilon_r - 1}} = 0.0825$  we get

$$Z_s = 88.25 \Omega$$

and  $\varepsilon_{co} = 5.711$

Similarly calculations for microstrip related parameters

Corresponding to  $\frac{w}{h} = 1.292$  under quasi static approximations and assuming the strip thickness to be negligible the expression for  $Z_M$  is given and the accuracy of this expression is reported to be better than 0.2% for  $\varepsilon_r \leq 128$  and  $0.01 \leq \frac{w}{h} \leq 100$

$$\varepsilon = \left\{ \frac{(\varepsilon + 1)}{2} \right\} + \left\{ \frac{(\varepsilon - 1)}{2} \right\} \left\{ 1 + \left( \frac{10}{2} h \right) \right\}^{a1} \quad 3.4(c)$$

Where

$$a = \left[ 1 + \left( \frac{1}{49} \right) \ln \left\{ \frac{\left( \frac{w}{h} \right)^4 + \left( \frac{w}{52h} \right)}{\left( \frac{w}{h} \right)^4 + 0.432} \right\} + \left( \frac{1}{18.7} \right) \ln \left\{ 1 + \frac{w}{18.1h^3} \right\} \right] \quad 3.4(d)$$

and

$$b = 0.564 \left( \frac{\varepsilon - 0.9}{\varepsilon + 3} \right)^{0.053} \quad 3.4(e)$$

So after substituting the known parameters in the equation we get

$$\varepsilon_c = 6.944$$

$$Z_M = \left\{ \frac{60}{\sqrt{\varepsilon}} \right\} \ln \left[ \left( \frac{h}{w} \right) F \left( \frac{w}{h} \right) + \left\{ 1 + \left( \frac{2h}{w} \right)^2 \right\}^{\frac{1}{2}} \right] \quad 3.4(f)$$

Where

$$F \left( \frac{w}{h} \right) = \left[ 6 + (2\pi - 6) \exp \left\{ - \left( 30.666 \frac{h}{w} \right)^{0.7528} \right\} \right] \quad 3.4(g)$$

So the value of  $Z_M = 43.36 \Omega$

$$Z_0 = \sqrt{Z_M Z_s} = 61.85 \Omega$$

For center frequency

$$\frac{\lambda_s}{\lambda_0} = \frac{1}{\sqrt{\varepsilon_{eff}}} \quad \lambda_s \text{ From coupler structure } = 4 \times \text{length from } T_1 \text{ plane to } T_2 \text{ plane} = 22.48$$

mm

Putting the numerical values of the parameters and calculating we get

$$f_c = 6.7404 \text{ GHz}$$

The increased characteristic impedance required for compensation

$$Z_s = \frac{Z_s}{2} \left[ \cot\left(\frac{\pi}{4} \sqrt{\frac{\epsilon}{\epsilon}}\right) + \tan\left(\frac{\pi}{4} \sqrt{\frac{\epsilon}{\epsilon}}\right) \right] \quad 3.4(h)$$

$$= 89.35 \Omega$$

The length of the compensation lines which is assumed to have approximate value is chosen according to the following expression

$$l_s = l \sqrt{\frac{\epsilon}{\epsilon}} \frac{2}{\pi} \arctan \left\{ \frac{Z_s}{Z_s} \left[ 1 - 2 \sin^2 \left( \frac{\pi}{4} \sqrt{\frac{\epsilon}{\epsilon}} \right) \right] \right\} \quad 3.4(i)$$

$$l_s = 0.58 \text{ mm}$$

The input susceptance obtained at the frequency  $f_c = 6.74024 \text{ GHz}$

$$B_c = \frac{1}{Z_s} \tan \left( \frac{2\pi l_s f \sqrt{\epsilon}}{c} \right)$$

$$B_c = 3.632 \times 10^3 \text{ mho}$$

So computed data which are relevant to realized coupler are as follows

Dielectric constant of the substrate ( RT duroid )  $\epsilon_r = 10.2$

Thickness of the substrate  $h = 0.325 \text{ mm}$

Width of the microstrip line  $w = 0.42 \text{ mm}$

Width of the slot line  $s = 0.14 \text{ mm}$

Width of the feed lines  $w_p = 0.62 \text{ mm}$

Length of the microstrip line  $l = 5.62 \text{ mm}$

Compensation length of the slot  $l_s = 0.58 \text{ mm}$

Effective permittivity of the microstrip medium  $\epsilon_{ee} = 6.944$

Effective permittivity of the slot line medium  $\epsilon_{eo} = 5.711$

Slotline characteristic impedance  $Z_s = 88.25 \Omega$

Microstrip line characteristic impedance  $Z_M = 43.36 \Omega$

Characteristic impedance of the coupler  $Z_O = 61.85 \Omega$

Effective permittivity of the coupled medium  $\epsilon_{eff} = 6.32$

Center frequency  $f_c = 6.7404 \text{ GHz}$

Input susceptance corresponding to center frequency  $B_c = 3.6 \times 10^{-3}$  siemens

Since  $B_c$  is shunted across the ends of the coupling section. So the reflection coefficients for odd mode excitation can be computed from the input susceptance of the transmission lines shunted with the susceptance  $2B_c$

$$\frac{\theta}{2} = \omega l \sqrt{\frac{\epsilon}{c}} = 1.0886 \text{ radian}$$

$$\frac{\theta}{2} = \omega l \sqrt{\frac{\epsilon}{c}} = 1.204 \text{ radian}$$

By putting the values of the parameters and calculating we get the input reflection coefficients for the compensated coupler as follows

$$\Gamma_{in} = \exp \left\{ -j2 \arctan \left[ \frac{2Z}{Z_0} \tan \left( \frac{\theta}{2} \right) + 2B_c Z \right] \right\} \Rightarrow \exp(j2.3873) \quad 3.4(j)$$

$$\Gamma_{out} = \exp \left\{ -j \arctan \left[ -\frac{2Z}{Z_0} \cot \left( \frac{\theta}{2} \right) + 2B_c Z \right] \right\} \Rightarrow \exp(j0.4534) \quad 3.4(k)$$

$$\Gamma_{in} = \exp \left\{ -j2 \arctan \left[ \frac{Z}{2Z_M} \tan \left( \frac{\theta}{2} \right) \right] \right\} \Rightarrow \exp(j1.956) \quad 3.4(l)$$

$$\Gamma_{out} = \exp \left\{ j2 \arctan \left[ \frac{Z}{2Z_M} \cot \left( \frac{\theta}{2} \right) \right] \right\} \Rightarrow \exp(j.4410) \quad 3.4(m)$$

For these input reflection coefficients we get

$$\text{Return loss} = 20 \log |S_{11}| \Rightarrow 9.74 \text{ dB} \quad 3.4(n)$$

$$\text{Coupled port output} = 20 \log |S_{31}| \Rightarrow 5.31 \text{ dB} \quad 3.4(o)$$

$$\text{Direct port output} = 20 \log |S_{21}| \Rightarrow 2.36 \text{ dB} \quad 3.4(p)$$

$$\text{Isolated port output} = 20 \log |S_{41}| \Rightarrow 19.6 \text{ dB} \quad 3.4(q)$$

$$\text{Directivity} = 20 \log \left| \frac{S_{41}}{S_{31}} \right| \Rightarrow 14.29 \text{ dB} \quad 3.4(r)$$

For the second question three definitions were investigated for  $Z_s$  and  $Z_M$ .  $Z^{(PV)} = V^2/(2P)$ ,  $Z^{(VI)} = V/I$  and  $Z^{(PI)} = 2P/I^2$  where  $P$  denotes the power transported in the direction of propagation of the mode.  $I$  denotes the longitudinal



current on one of the two electrodes of the slot line or the strip conductor of the microstrip transmission line and  $V$  denotes the voltage over the slot in the slot line or the voltage between the center of the strip conductor of the microstrip line and the ground. The conversion formula is

$$Z^{(LI)} = [Z^{(VI)}]^2 / Z^{(PV)}$$

Linear interpolation is possible with good approximation (error < 0.5 percent) between 4 GHz and 6 GHz and between 6 GHz and 8 GHz.

We first consider the 3 dB coupler which is matched to  $Z_0 = 50 \Omega$ . The four positions of the reference planes  $T_1$  and  $T_2$  were used to calculate the theoretical values of  $a_{21}(f)$  and  $a_{31}(f)$ . It can be seen that the closest agreement with the measured frequency responses occurs for the reference plane positions defined by

$$d = \frac{w_l}{2} \quad \text{which can be alternatively expressed by } l = a + \frac{w_p}{2}.$$

Thus  $T_1$  and  $T_2$  are located approximately  $w_p/2$  within the outer edges of the feeding lines of width  $W_p$ . Larger value of  $l$  result in frequency responses that deviate more sharply from the measurements.

The theoretical responses for  $a_{21}(f)$  and  $a_{31}(f)$  were calculated using the one  $Z_s$  definition, the position  $d = \frac{w_l}{2}$  of the reference planes. For the microstrip mode we used the definition  $Z_M$ . The difference between the  $Z_M$  definitions is below 5 percent and in the first approximation negligible. Also negligible is the terminal capacitance of the slot line open circuits. When  $Z_s$  definition is used, the theoretical responses for  $a_{21}(f)$  should almost coincide with the measured response of the coupling section without feeding lines, which is obtained by subtracting the 0.1 dB attenuation to the feeding lines. In our case there is deviation because of alignment problem. As slot is not completely beneath the microstrip line on the other side of the substrate. Perfect alignment is only possible if alignment is done with mask aligner. Although theory should yield lower values as the dissipative losses are not being taken in to account. The computed values of  $a_{21}(f)$  and  $a_{31}(f)$  are average of 0.4 dB below the measured

values. This difference is attributed due to the ohmic, dielectric and the radiation losses at discontinuities that are not taken in to account in the theory. Thus the definition of  $Z_S$  is the more preferable one for the coupler design.

The choice of the reference plane positions and the slot line impedance definition obtained for the matched coupler is also applies to coupler having slot width  $s=0.14\text{mm}$ .

As the accuracy of the curves concerns, the following information should be added here. Measurement of the substrate thickness  $h$  over the whole substrate resulted in maximum deviation of  $8\mu\text{m}$  from the nominal value of  $0.325\text{mm}$  which keeps the change in  $Z_S$  and  $Z_M$  below 0.5 percent and that of the  $\epsilon_{eo}$  and  $\epsilon_{ec}$  0.1 percent. Also the dielectric constant tolerance  $\Delta\epsilon/\epsilon_r < 3\text{percent}$  results in a change of  $< 2\text{percent}$  for  $Z_S$ ,  $Z_M$ ,  $\epsilon_{eo}$  and  $\epsilon_{ec}$ . Thus the total maximum error for  $Z_S$  and  $Z_M$  of 2.5 percent is noticeably smaller than the typical difference of 8 to 20 percent between the various  $Z_S$  definitions. Similarly the possible frequency shift of  $< 1\text{percent}$  for the theoretical results due to tolerances in  $\epsilon_r$  and  $h$  is noticeable smaller than the center frequency shift of 8 to 30 percent caused by different reference plane positions  $d=0$  to  $1.5 w_p$ . The measurements were performed using PC interfaced power meter, directional coupler and phase locked frequency synthesizer.

### **3.5 Measurements and results of the microstrip-slot coupler**

Measurements are done by network analyzer made out by interconnecting power meter, direction coupler and frequency synthesizer (2-20 GHz) and its operation controlled by computer. Only the magnitude of  $S_{11}$ ,  $S_{21}$ ,  $S_{31}$  and  $S_{41}$  of the device are characterized.  $S_{11}$ ,  $S_{21}$ ,  $S_{31}$  and  $S_{41}$  represent the reflection, Transmission, coupling and isolation parameters respectively. Figure 2.1 illustrates the microstrip slot coupler. It can be seen that the measured center frequency is in good approximation with the theoretical value 6.70404 GHz. The combined plot of experimental results taken is given in figure 3.3.

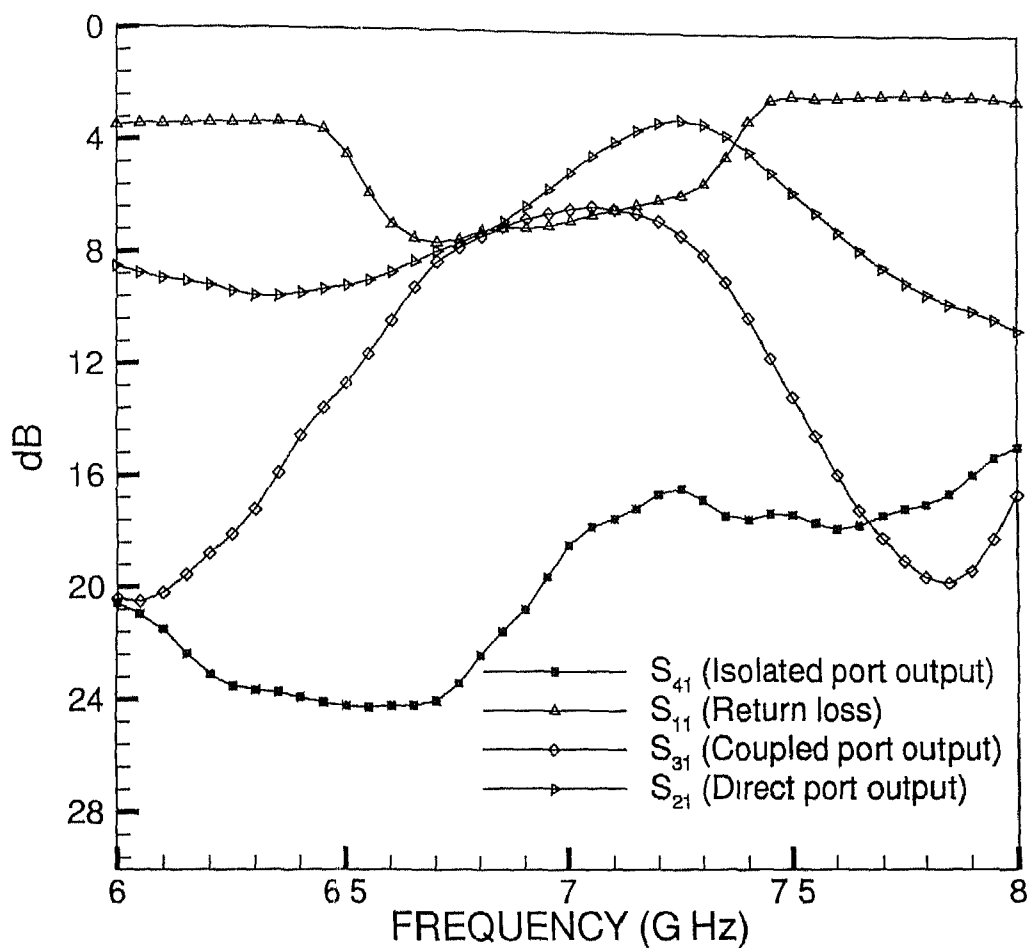


Figure 3.3 Plot showing measured magnitude of  $S_{41}$ ,  $S_{11}$ ,  $S_{31}$  and  $S_{21}$  of microstrip slot coupler

#### Specifications of coupler

Coupling = 6 dB

Return loss = 7.5 dB

Coupling band width = 15%

Isolation band width = Better than 15%

Isolation = Better than -20 dB

Center frequency = 7.0 GHz

The isolation is better than the theoretical value 19.6 dB but direct port output 3.5 dB is fairly in good agreement over the limited range of frequencies with the

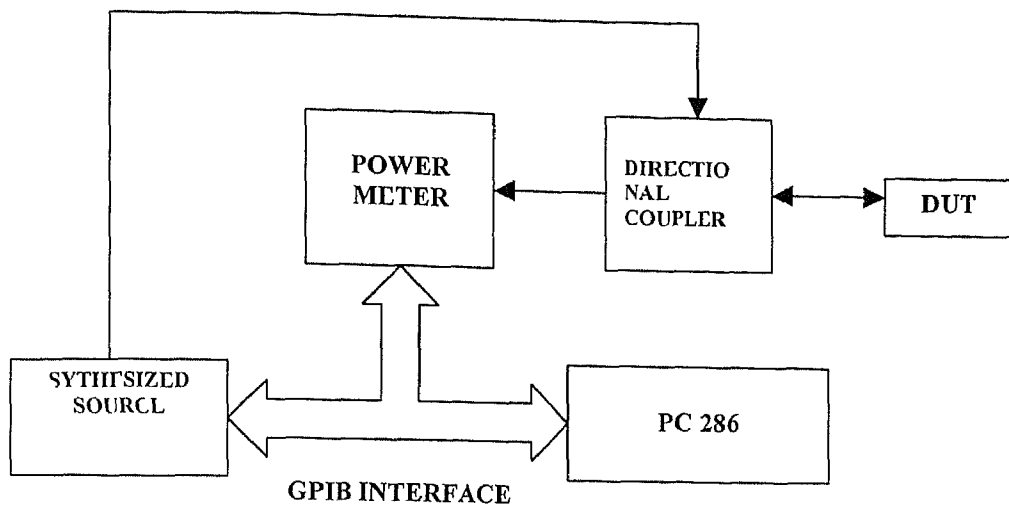


Fig 3 4 Block diagram of the experimental setup for return loss measurement

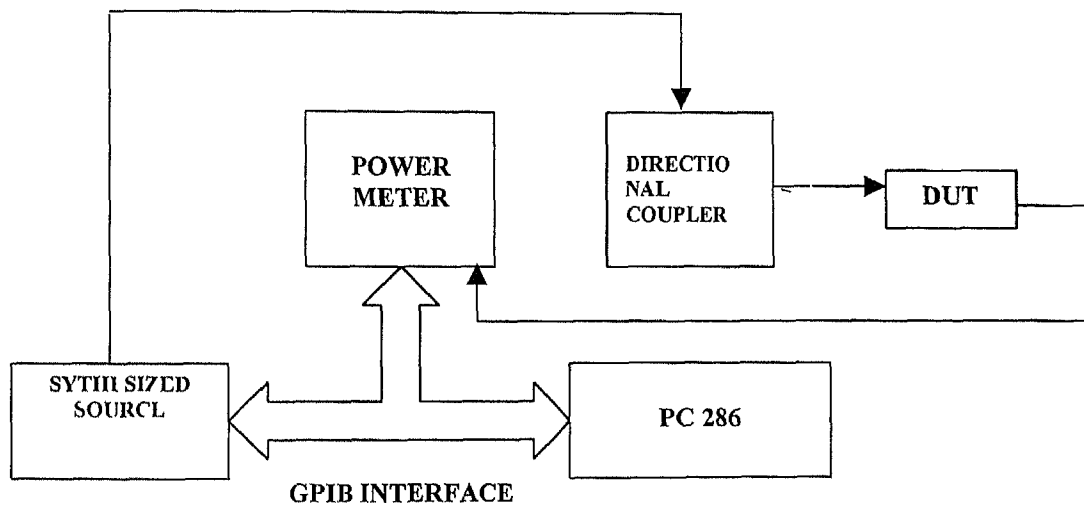


Fig 3 5 Block diagram of the experimental setup for direct port coupled port and isolated port output

The isolation is better than the theoretical value 19.6 dB but direct port output 3.5 dB is fully in good agreement over the limited range of frequencies with the theoretical predicted value 2.36 dB. Coupling from 6.5 GHz – 7.5 GHz is in the range 6dB to 7 dB which is not in good agreement with the approximate theoretical predicted value 5.31 dB. It can also be seen from the plot that return loss is 7.5 dB which is not in good agreement with the theoretical predicted value 9.74 dB in the band of interest. This is due to mismatch and transition losses at connection ports.

# CHAPTER 4

## Summary and conclusion

### 4.1 Summary

A novel method of analysing microstrip slot coupler theory and realization of the same had been described in this thesis. The microstrip slot coupler is special classes of couplers for that type of coupler expression of scattering parameters are derived in terms input reflection coefficients by using even odd mode analysis of a reciprocal passive linear four port network with double symmetry with reference to the two symmetry planes. Derivation of the scattering parameters is done in both the cases i.e. without compensation and with compensation lines. Compensation is accomplished by lengthening the slot line to capacitive loading at the ends of the coupled region. The isolation has been improved by 5-10 dB as compared to the uncompensated structure. The frequency response of the isolation depends on the ratio of  $\frac{l_s}{Z_0}$ . Special cases of microstrip slot couplers retaining ideal theoretical approach

can not be realized with the real coupler configuration because in this case  $\epsilon_r < \epsilon_{ee}$  however it can be used to derive design equations for the real couplers. The ideal coupler is characterized by the reflection coefficient  $S_{11}$  being zero and  $S_{41}$  being zero i.e. infinite directivity at all frequencies. For real uncompensated commercial coupler of conventional dimension  $l = 0$ ,  $\epsilon_{ee} > \epsilon_{eo}$  and  $v_o > v_e$ . The real couplers with compensation ( $\epsilon_{eo} < \epsilon_{ee}$ ,  $l > 0$ ) and matching i.e.  $S_{41}=0$ ,  $S_{11}=0$  can be realized with the aid of the compensation lines only at a single arbitrary chosen frequency viz. the compensation frequency  $f_{co}$  for that it is essential that matching condition should be satisfied.

Practical designing aspects of the microstrip slot coupler on RT/Duroid substrate with compensation are treated in the supplementation of the theoretical analysis. Emphasis is given to the realization of 4.5 dB coupler on the 0.325 mm thick

substrate with  $\epsilon_r = 10.2$  substrate. We had design data for the realization 4.5 dB coupler on the  $\text{Al}_2\text{O}_3$  substrate with relative permittivity of 9.8 and substrate thickness 0.635 mm. But due to non availability of the compatible substrate the fabrication is done on the substrate having thickness 0.325 mm for the slotline 0.14 mm on the ground plane and microstrip width 0.42 mm on the other plane for the mismatch condition. The theoretical coupling is 5.31 dB for the given substrate having same dimensions of slot line and microstrip as was implemented on the substrate of thickness 0.625 mm. The coupler is working satisfactory with center frequency 7.0 GHz. The practical coupling value is not equal to the approximately calculated value because of the slot is not accurately beneath the microstrip in the ground plane this is because of the non availability of the mask aligning facility. If the alignment is done with the mask aligner we can get coupler of which coupling has close agreement with the calculated values for the perfect matching condition. For the isolation improvement compensation is added in this type of coupler which has removed the junction effect with enhancing the directivity. In uncompensated type of coupler such a good directivity can not be obtained. In this thesis comparison has been made for the theoretical and practical parameters.

## 4.2 Conclusion

Scattering parameters of the coupler are derived in this thesis by using the even odd mode analysis for four port network having double symmetry. Equations have been derived for the frequency dependent scattering parameters  $S_{ij}$  of the microstrip slot coupler whereby different phase velocities  $v_e, v_o$  of the even and odd modes as well as the mismatch are included through the arbitrary choice of the characteristic impedance  $Z_0$  and the even mode and odd mode characteristic impedances  $Z_e, Z_o$ . The analysis was extended to cover the coupler with supplementary slot lines as conventionally used for compensation. For  $v_e = v_o$  the parameters of an ideal TEM coupler are realized. Simple design specifications are derived for the compensation slot lines.

Comparison of the measurement of coupler with a slot width 0.14 mm realized on RT/Duroid substrate with results yielded by the theoretical analysis show close agreement under the following condition. First the reference planes for the ends of the coupling section must lie within about half width of the feeding lines with center frequency in the region of 7 GHz. Second the characteristic impedance  $Z_0$  defined by the current and voltage must be used for the slot line. Third in order to take of the transmission line loss the theoretically calculated parameters must be displaced by a fixed value – here approximately 0.4 dB – towards higher attenuations. Estimates have shown the theories of the standard microstrip transmission line and the standard slot line to be sufficient for the practical designing of the coupler design.

### 4.3 Scope for future work

The microstrip slot line coupler can be integrated on the high relative permittivity substrate to improve the dispersion loss of the coupler as effective permittivity in slotline propagation mode becomes much lower in comparison to the relative permittivity of the used substrate. If available the integration of the coupler can be done on the transparent substrate having high relative permittivity of the order 20 to 30 which will eliminate the error occurring in the alignment of the microstripline and slotline. If the transparent substrate is not available the alignment requires mask alignment technique during fabrication process. For accurate theoretical calculation of parameters full wave analysis can be used with which we can get design parameters very close to practical values of the implemented coupler.



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